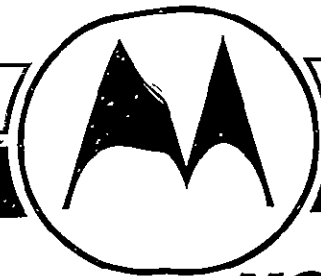
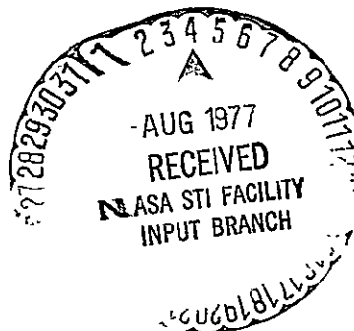


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MOTOROLA INC.
Government Electronics Division



MONOPULSE RECEIVER FINAL REPORT

April 1977
In Response to
Contract No. JPL 954588

Approved by:

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Aerospace Rf Section

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ABSTRACT
MONOPULSE RECEIVER
DEVELOPMENT AND DEMONSTRATION

This document is the final report on the Monopulse Receiver Program performed by Motorola, Inc., Government Electronics Division (GED) for the Jet Propulsion Laboratory under JPL contract No. 954588.

Two dual-channel systems were analyzed, designed, and tested. The two systems investigated were Time Division Multiplexing and Quadrature Combining.

Analyses performed include the following:

- Boresight error as a function of the error channel bandwidth
- Error channel interference with the sum channel functions
- Threshold performance
- Error channel crosstalk, linearity and drift as a function of signal level, Doppler and environment

Test results indicate that the Time Division Multiplexing System meets the design goals of the program. However, careful selection and alignment of all gain-controlled amplifiers are required along with temperature compensation of the angle channel gain.

The Quadrature system crosstalk performance is comparatively poor (-15 dB) with respect to the -30 dB requirements and this consequently affects gain tracking performance. Gain tracking was found to be ± 20 percent rather than the ± 5 percent specification.

Data for the two systems is compared and recommendations presented in this report.

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SECTION I

1. SUMMARY

This document is the final report on the Monopulse Receiver Program performed by Motorola, Inc., Government Electronics Division (GED) for the Jet Propulsion Laboratory under JPL contract No. 954588. Two dual-channel systems were analyzed, designed, and tested. The two systems investigated were Time Division Multiplexing and Quadrature Combining.

Analyses performed include the following:

- Boresight error as a function of the error channel bandwidth
- Error channel interference with the sum channel functions
- Threshold performance
- Error channel crosstalk, linearity and drift as a function of signal level, Doppler and environment

Test results indicate that the Time Division Multiplex System meets the design goals of the program. However, careful selection and alignment of all gain-controlled amplifiers is required along with temperature compensation of the angle channel gain.

The Quadrature system crosstalk performance is comparatively poor (-15 dB) with respect to the -30 dB requirements and this consequently affects gain tracking performance. Gain tracking was found to be ± 20 percent rather than the ± 5 percent specification.

Data for the two systems is compared and recommendations are presented herein.

SECTION 2

2. INTRODUCTION

The intent of the program was to design a monopulse error channel capable of meeting the performance requirements defined in Section 3 and compatible with the NASA Standard Near Earth and Deep Space Transponders. The dual-channel monopulse system concept was the result of the earlier study and tradeoff analysis conducted on the multimission transponder study under Contract No. 957975. The Sum Channel Receiver, supplied by JPL, was the Microminiature Receiver which approximates the design of the Standard Transponder, with the primary differences being slight variations in gain distribution and intermediate frequency bandwidths. Neither of these variations are of a magnitude to obviate actual measured performance with that predicted for a standard design. The error channel RF converter, first IF and second IF modules were fabricated using beam lead technology. All other modules were breadboarded utilizing discrete components but with the capability of being fabricated with beam lead components which is compatible with the standard transponder design.

SECTION 3

3. PERFORMANCE REQUIREMENTS

The following requirements were placed on the Monopulse Receiver error channels.

- **NST Compatibility**

The Monopulse Receiver design shall be electrically compatible with the NE-NST described in STD-336-M01-DS.

- **Sum Channel Interference**

1. The error channel shall not degrade the sum channel threshold by more than one-half dB or the receiver best lock frequency by more than ± 1 ppm at $+26^{\circ}\text{C} \pm 2^{\circ}\text{C}$.
2. The error channel shall produce no signals that will interfere with command or ranging functions of the sum channel.

- **Monopulse Characteristics**

1. The signal-to-noise (S/N) ratio in the error channel output shall be greater than +6 dB measured in a $B_{Lo}/10$ low pass noise bandwidth when the sum channel carrier power is 10 dB or greater above carrier threshold (carrier threshold is defined as zero dB S/N in the sum channel carrier loop) and the error channel signal is 10 dB below the sum channel carrier power.
2. The effect of a signal in one error channel on the output of the other error channel shall be less than that of a 30 dB weaker signal in the other error channel.
3. The variation in error signal output at a fixed error angle shall be less than 5 percent of the maximum output as the input levels varies from -50 dBm to threshold and as the temperature varies over the FA range specified in the Environmental Requirements paragraph.
4. The maximum error signal out of either channel with no error input shall be no greater than 2 percent of the output that occurs when the error channel signal is equal to the sum channel signal.

- **Dynamic Requirements**

The Monopulse Receiver shall track properly with slew rates of 30 dB/s or less in the error channel signals.

- **Environmental Requirements**

1. The receiver shall be designed to operate over the environmental conditions of STD-336-M03-ER.
2. The demonstration breadboard need only be tested from -10°C to $+55^{\circ}\text{C}$.

- **Physical Constraints**

The unit shall be in clean breadboard form; size and weight are not critical. Standard or M-series type hardware shall be utilized where cost is not increased or where performance is not compromised by using alternate circuitry.

- EMI Requirements

The receiver shall be designed with good engineering standards and shall operate without degradation in the electrical environment expected on an antenna range or laboratory.

- Reliability

1. Good engineering practice shall be followed.
2. No detailed stress analysis or reliability data shall be required.

SECTION 4

4. SYSTEM DESCRIPTION

The Monopulse Receiver consists of a sum channel and a single error channel. The sum channel provides the phase locked reference signals and the automatic gain control (AGC) voltage to the error channel. The two S-Band error signals are combined either by time division multiplexing (TDM) or by quadraphasing onto a single error channel of identical signal path design to the sum channel receiver after all gain and phase corrections. The error signals are demultiplexed and demodulated to provide two dc steering signals to the antenna pointing control system. The sum channel utilized on this program was the Microminiature Receiver developed for JPL on a previous contract. This receiver was modified to provide the proper reference signals for the angle channels. Figure 4-1 is a block diagram of the sum channel with modifications. The Microminiature X8 LO Multiplier (X2, X2, X2) was replaced with a higher output (shock recovery diode, X8) module. A 90-degree hybrid coupler was added to the output to provide the error channel first LO signal at a power level of +5 dBm.

The second LO and detector reference signals were resistively tapped and supplied to the error channel as low level references, (second LO: -22 dBm; F2 reference: -48 dBm). The sum channel AGC voltage was also resistively tapped and supplied to the error channel IF amplifiers. In addition, the two beam lead submodules containing the IF amplifiers with AGC, in the sum channel were replaced using devices which were gain matched to the similar error channel units. This was necessary to provide the optimum gain tracking performance with signal level and temperature variations.

The primary differences between the Microminiature Receiver and the Standard Transponder Receiver with which the error channels would ultimately perform are the addition of an S-Band preamplifier and the replacement of the 12 kHz crystal filter with a 250 kHz L-C filter. Neither difference would affect the ultimate monopulse system performance.

4.1 TDM System

The TDM System is shown in Figure 4-2. A single-pole double-throw pin diode switch is used to sample, at a 50 percent duty cycle, the two error signals. The switching frequency is approximately 50 Hz. The switch provides at least 40 dB of isolation. The combined signals are then processed through a preselector and mixer preamplifier which would ultimately be identical to the sum channel units. The mixer preamplifier design in the breadboard error channels is slightly different in design, which affects the gain and phase tracking capability of the breadboard unit to a slight degree. The first and second IF modules are identical through the signal path. In the error channel first IF, the AGC loop filter amplifier and ranging channel amplifiers were deleted and replaced with a second LO buffer amplifier and AGC gain and slope adjust amplifiers. In the second IF the circuitry associated with the loop phase detector was omitted. Both IF modules were fabricated using beam lead submodules in the signal path which were identical to the sum channel units. The LO and AGC circuitry were built using discrete devices which are capable of being built in beam lead form. The detector module contains separate circuitry for each of the error channels. The IF signals are

processed through a temperature compensated amplifier, adjusted to compensate gain differences between error and sum channels. An identical device is then used to select the proper time-multiplexed signal. The phase detector and operational amplifier then provide the basic 50 Hz sampled signal to the sample and hold circuit and operational amplifier which produces a dc output voltage proportional to the amplitude or phase of the received signals. The module also contains the buffer amplifiers and phasing adjustments necessary to condition the detector reference signals.

All detector/demultiplexer circuitry was built using discrete components; however, all but the sample and hold and phasing adjustments could be built in beam lead form.

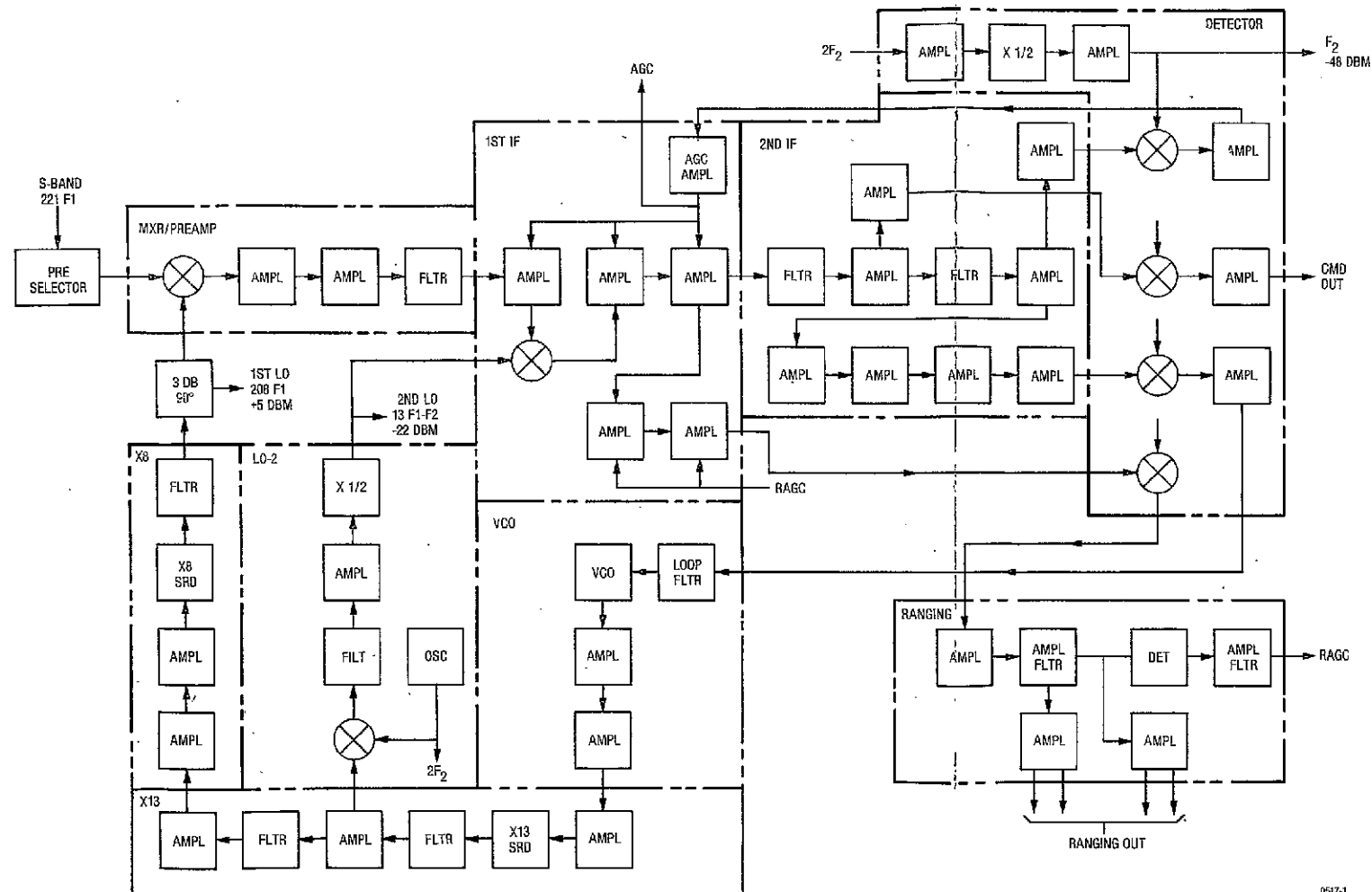
The timing module contains the basic clock, frequency dividers and driver circuitry necessary to provide a symmetrical fast rise time 50-Hz signal for the multiplexing devices.

The timing and sample and hold circuitry required ± 12 volts dc which is not available in the Microminiature Receiver. Consequently, laboratory supplies were used to provide these functions. The Standard Transponder, however, does contain ± 12 volts dc capability to power these circuits.

4.2 Quadrature System

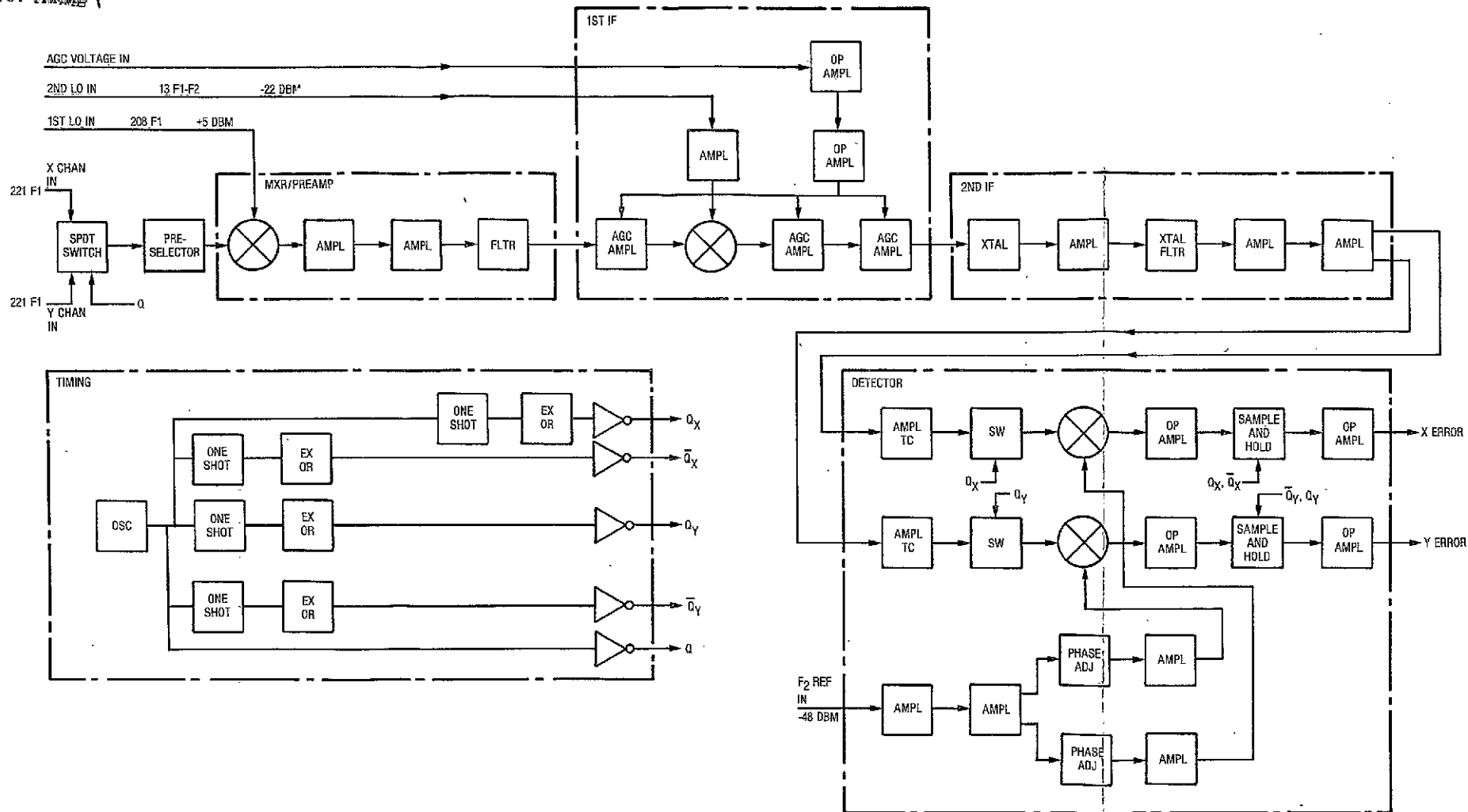
The quadrature system shown in Figure 4-3 utilizes the same RF converter, first IF and second IF modules as the TDM System. A 90-degree hybrid coupler is used to combine the two error signals in quadrature. The IF signals are then processed in quadrature detectors in order to separate them. The switched RF amplifier, detector, and operational amplifier form a chopper-stabilized system which provides the proper dc balance performance. The timing signals for this function are provided by the quadrature timing module.

FOLDOUT FRAME 1



0617-1

Figure 4-1. Microminiature Receiver Block Diagram



0517-3

Figure 4-2. TDM Error Channel Block Diagram

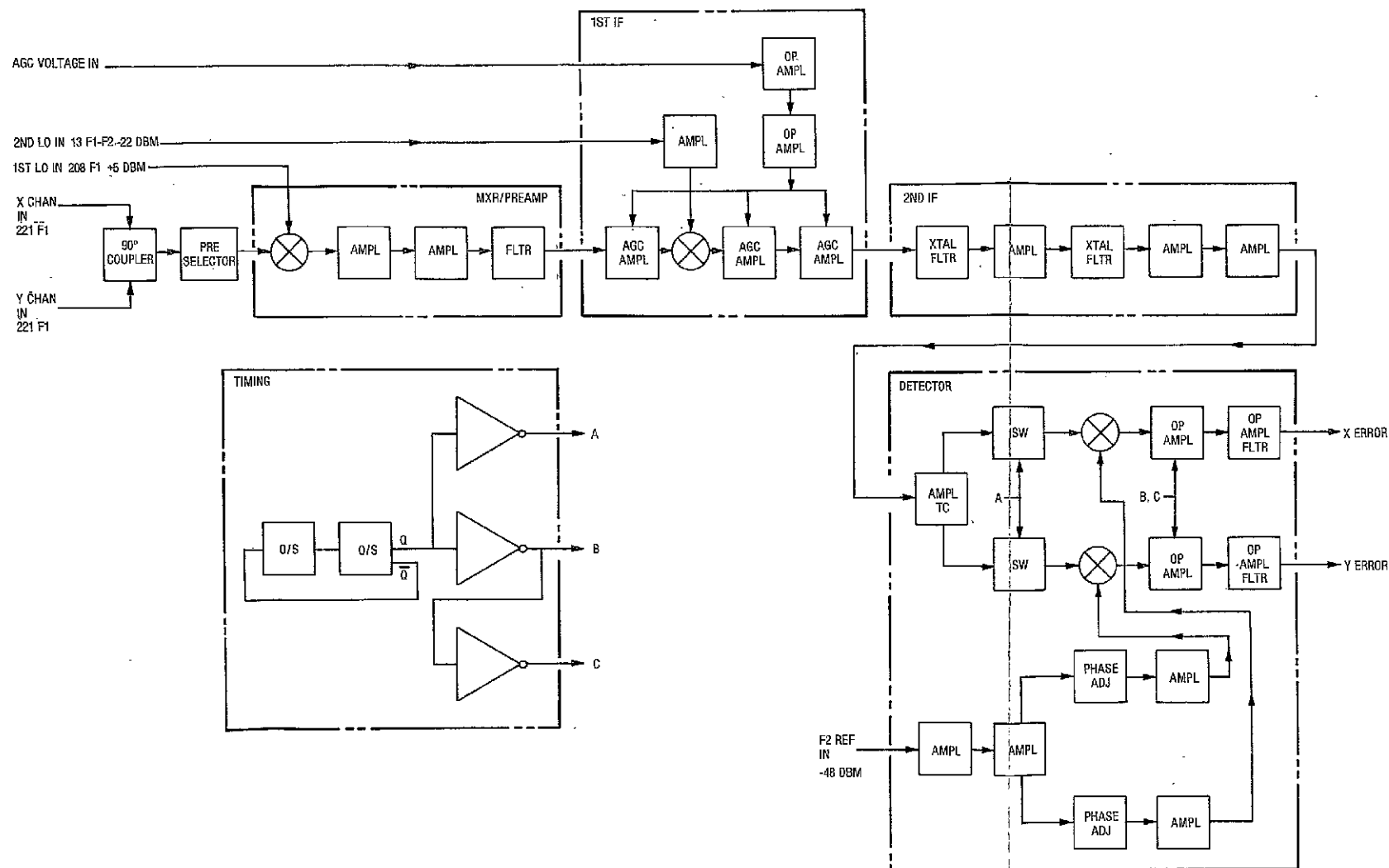


Figure 4-3. Quadraphase Error Channel Block Diagram

0517-2

SECTION 5

5. ANALYSIS

The following system analysis was based on the Standard Transponder, while the tests were performed using the Microminiature Transponder. The Standard Transponder has a noise figure of 5.5 dB and IF bandwidth of 250 kHz, while the Microminiature has a 7.5 dB noise figure and IF bandwidth 20 kHz. Both transponders have identical predetection noise bandwidth of 4 kHz. System analysis was performed to evaluate the following parameters:

- Boresight errors as a function of the error channel bandwidth using typical imbalances and aging factors
- Error channel interference with the sum channel command, ranging, and tracking functions
- Threshold performance as a function of the error channel bandwidth
- Error channel crosstalk, linearity, and drifts and behavior under signal dynamics and temperature variations

The first three items above are identical for the two system studies while the last item is directly related to the particular system under discussion. Figure 5-1 depicts the monopulse clock parameters.

5.1 Time Division Multiplexing System

5.1.1 BORESIGHT ERRORS AS A FUNCTION OF ERROR CHANNEL BANDWIDTH

Boresight shift normalized to the antenna half-power beamwidth, assuming a Gaussian response (Figure 5-2) is:

$$\frac{\theta_o}{\theta_3} = \frac{\theta_3}{\theta_m} (0.180) \ln \left[\left| \frac{A_2}{A_1} \right| \left(-\sin \alpha \tan \beta + \sqrt{1 + \sin^2 \alpha \tan^2 \beta} \right) \right]^*$$

where

- θ_o = Boresight error
- θ_3 = 3 dB beamwidth
- θ_m = Squint angle
- A_2/A_1 = Precomparator amplitude unbalance
- α = Precomparator phase shift between channels
- β = Post-comparator phase shift between channels

When,

$$\frac{A_2}{A_1} = \alpha = \beta = 0 \text{ or } \left| \frac{A_2}{A_1} \right| = \alpha = 0 \text{ and } \beta \neq 0$$

$$\frac{\theta_o}{\theta_3} = 0 = \text{No boresight error}$$

* Dual S- and Ku-Band tracking feed for a TDRS Reflector Antenna, Martin Marietta Aerospace Corp, No. OR 13225, Aug 1974 Prepared for Goddard Space Flight Center.

The receiver only adds to boresight shift through β and if no precomparator errors exist the receiver does not contribute any error. Thus, in order to evaluate receiver contributions, arbitrary values of $A_2/A_1 = (1.05/0.95)$ (0.4 dB) and $\alpha = 3$ degrees will be assumed. Also assume a 1-dB crossover, $\theta_3/\theta_m = 3.46$.

With these assumptions,

$$\frac{\theta_o}{\theta_3} = (0.623) \ln \left[\begin{array}{c} 1.05(-\sin 3 \tan \beta + \sqrt{1 + \sin^2 3 \tan^2 \beta}) \\ \text{or} \\ 0.95 \end{array} \right] \quad (5-1)$$

The receiver predetection bandwidth is set by a narrowband crystal filter in the second IF module. Its bandwidth is 3000 Hz for the Standard Transponder. The tracking of the filters frequency shift with age will provide the primary phase shift (β) of the angle channel.

Crystal aging data from the standard IF filter crystal and from other sources indicates an average frequency shift of 0.75 ppm at 1 year with a 1-ppm tracking maximum between crystals. This could be reduced to approximately 0.25 ppm with crystal selection, however the 1-ppm figure will be used in the following calculations.

The phase shift of a single-pole bandpass filter is

$$\beta_1 = \tan^{-1} 2 \left(\frac{\omega_c - \omega_i}{\omega_{3dB}} \right)$$

where ω_c = Center frequency = 12.25 MHz

ω_{3dB} = 3 dB bandwidth

ω_i = Frequency of interest

The expected frequency shift is 0.75 ppm $\pm$ 0.5 ppm. The untracked phase shift will be:

$$\beta = \tan^{-1} \left(\frac{2 \times 10^{-6} \times 12.25 \times 10^6}{\Delta f_{3dB}} \right) = \tan^{-1} \left(\frac{24.5}{\Delta f_{3dB}} \right) \quad (5-2)$$

Substituting equation (5-2) into (5-1):

$$\frac{\theta_o}{\theta_3} = (0.623) \ln \left[\begin{array}{c} 1.05 \left(-\sin 3 \tan \tan^{-1} \left[\frac{24.5}{\Delta f_{3dB}} \right] \right. \\ \text{or} \\ 0.95 \end{array} \right. \\ \left. + \sqrt{1 + \sin^2 3 \tan^2 \tan^{-1} \left[\frac{24.5}{\Delta f_{3dB}} \right]} \right]$$

which reduces to:

$$\frac{\theta_o}{\theta_3} = (0.623) \ln \left\{ \begin{array}{c} 1.05 \\ \text{or} \\ 0.95 \end{array} \left[-\sin 3 \left(\frac{24.5}{\Delta f_{3dB}} \right) + \sqrt{1 + \sin^2 3 \left(\frac{24.5}{\Delta f_{3dB}} \right)^2} \right] \right\} \quad (5-3)$$

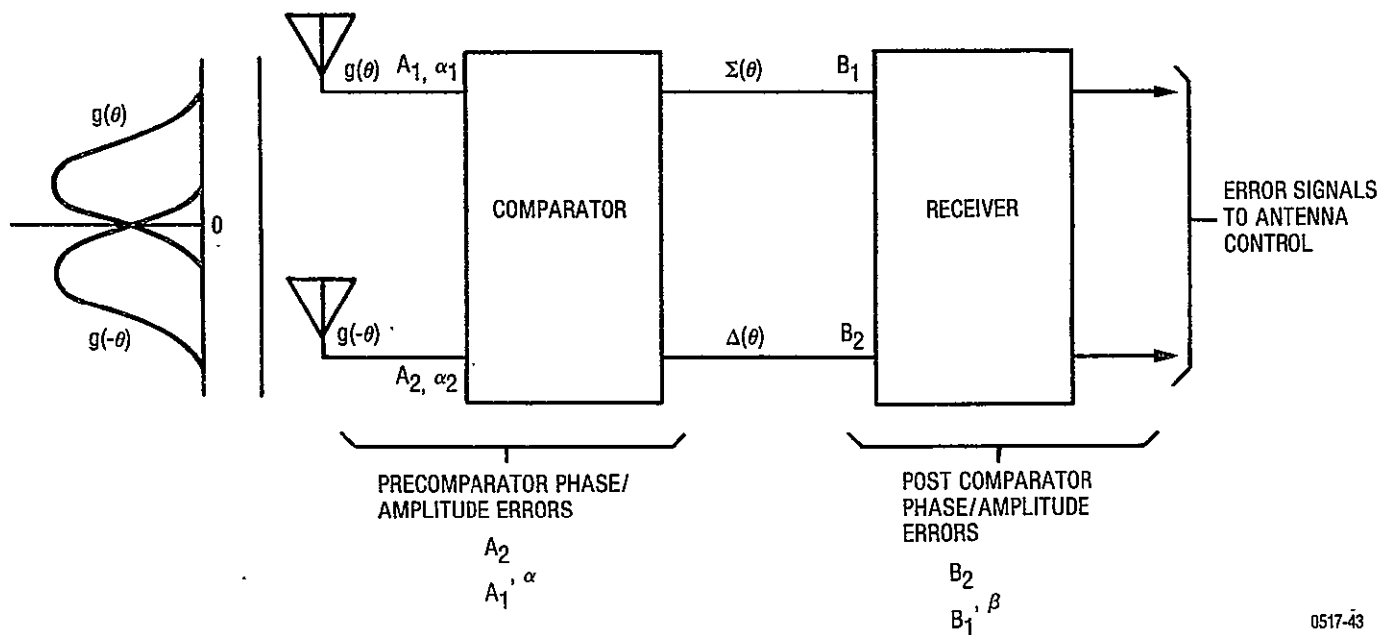


Figure 5-1. Monopulse System Block Diagram

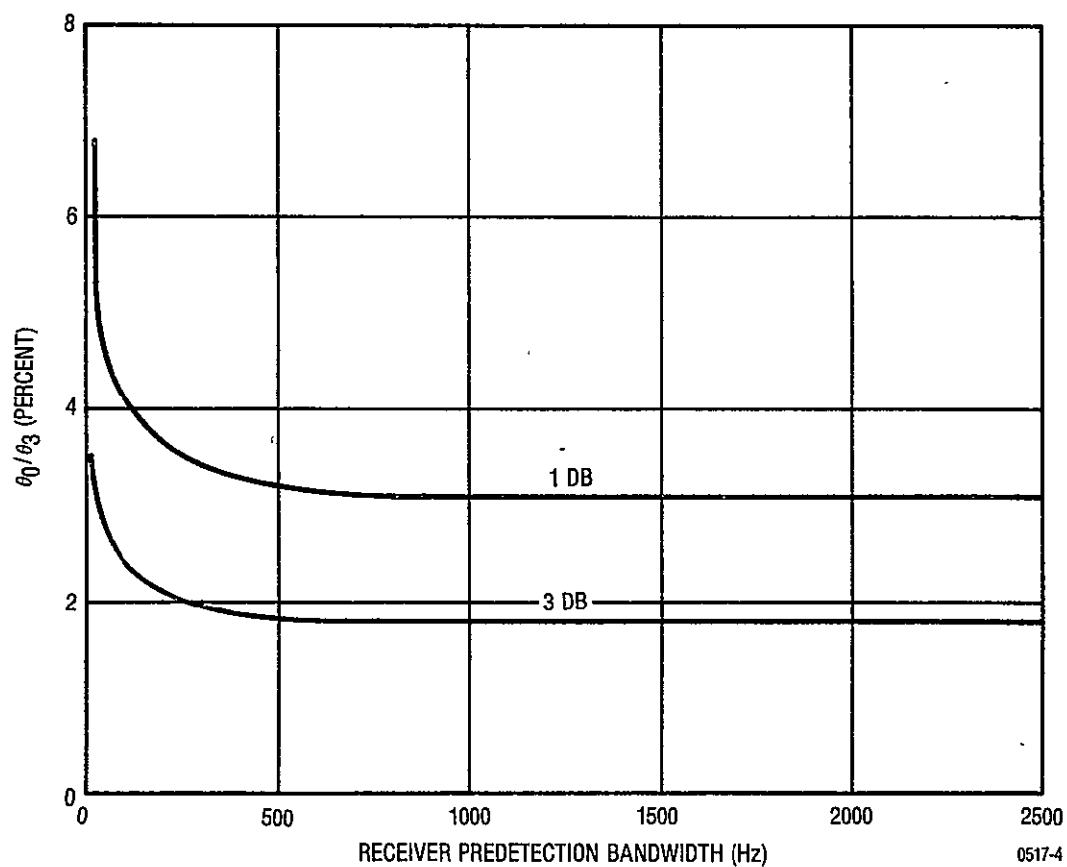


Figure 5-2. Boresight Error in Percent of 3 dB Beamwidth vs. Error Channel Bandwidth (Antenna Crossover 1 dB and 3 dB)

Figure 5-2 shows the boresight shift as a function of filter bandwidth for the stated conditions. It is seen that no significant error is contributed by the receiver for channel bandwidths greater than 1 kHz.

However, phase shifts due to other channel components will exceed those of the predetection filter with time and temperature variations. Figure 5-3 displays boresight error as a function of various pre- and post-comparator errors.

5.1.1.1 Boresight Angle Detector Gain

Both amplitude unbalance, $|B_1/B_2|$, and phase angle shifts, β , in the receiver will cause gain changes. The error is described by the equation:

$$A(\mu) = \left| \frac{B_1}{B_2} \right| \cos \beta$$

where $A(\mu) = 1$ for no error, i.e. $\mu = 0$

The receiver is specified to contribute a maximum of 5 percent variation in error channel output level over signal dynamics and temperature.

Therefore, $1.05 \geq A(\mu) \geq 0.95$

The estimate of error channel phase change with temperature, signal level, and Doppler is 10 degrees. Since phase compensation is difficult to achieve, this variation will be tolerated and the resultant maximum gain variations allowance computed.

$$\left| \frac{B_1}{B_2} \right| \leq \frac{A(\mu)}{\cos \beta}$$

$$\text{then } \left| \frac{B_1}{B_2} \right| \leq \frac{0.95}{\cos 10} \leq 0.9647$$

$$\text{and } 20 \log \left| \frac{B_1}{B_2} \right| = 0.3125 \text{ dB}$$

5.1.2 ERROR CHANNEL EFFECTS ON SUM CHANNEL

Since the error channel signals are processed in a separate channel, there is no direct interference. All effects are determined strictly by the isolation of the sum channel references supplied to the error channel and by the power lines.

The paths are:

- First LO
- Second LO
- F_2 reference
- AGC
- Power supply

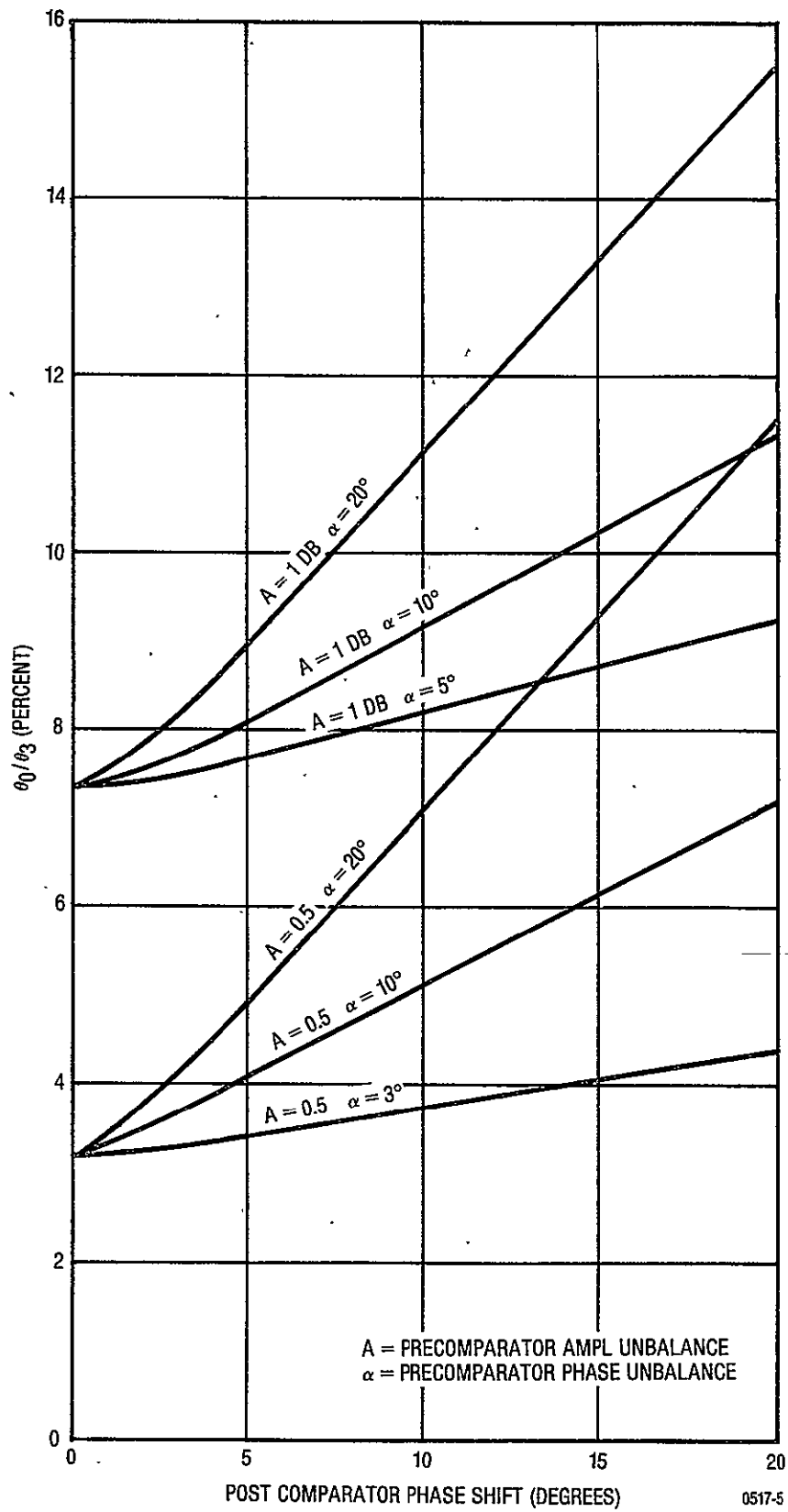


Figure 5-3. Boresight Error in Percent of 3 dB Beamwidth vs. Error Channel Phaseshift (1 dB Crossover)

5.1.2.1 Error Channel Leakage

Examination of these paths shows:

1. Error channel S-Band signals will leak to the sum channel mixer through the first LO line. Isolation is provided by:

First mixer signal to LO port	> 30 dB
3 dB hybrid output to output	> 30 dB
TOTAL	> 60 dB Isolation

No discernible effect.

2. Error channel IF signals will leak to the sum channel IF through the second LO line. Isolation is provided by:

1-2PH1 mixer	> 80 dB
1-2PH3 amplifier	> 40 dB
TOTAL	> 120 dB

No discernible effect.

3. Error channel IF signals will leak through the second MIXER to the sum channel IF. Isolation is provided by:

3-2PH3	> 120 dB
2-2PH1	> 80 dB
TOTAL	> 200 dB, or more realistically, 80 to 100 dB

No discernible effect.

4. Error channel signals leaking to sum channel through RF amplifiers and AGC amplifiers. Isolation is provided by:

1-2PH3	-40 dB
2-1741 op amplifiers	> 40 dB
TOTAL	> 80 dB

No discernible effect.

5. IF and chopper frequencies will ride power supply lines and affect the sum channel. The worst case will be the TDM chopper frequency since little filtering is provided at low frequencies.

One degree of sum channel phase error or 0.1 dB of amplitude interference will be permitted. The two most susceptible modules in the sum channel are the VCO and detector. The detector sensitivity is 67 millivolts detector output for 1 volt dc line change. The VCO sensitivity is 130 Hz/Vdc. Assume that the sensitivity is the same for 50 Hz signals.

The phase detector loop sensitivity is 8.3 mV/deg therefore for 1 degree of interference on the dc line:

$$\frac{8.3}{0.067} = 124 \text{ mV peak is allowable.}$$

The VCO dc line sensitivity is

$$130 \times 360 \text{ deg/V} = 46.8 \times 10^3 \text{ deg/V}$$

$$\frac{1}{46.8 \times 10^3} = 21.4 \text{ } \mu\text{V peak is allowable.}$$

The CAD loop sensitivity is 0.006 V/dB, and for 0.1 dB of dc line interference,

$$0.006 \text{ V/dB} \times 0.1 \text{ dB} \times \frac{1}{0.067} \text{ V} = 9 \text{ mV peak dc is allowable.}$$

From the above it is seen that the VCO constrains the allowable dc line interference and 21.4 microvolts peak is the maximum allowable level.

Using 20 microvolts peak as the maximum allowable chopper signal on the power lines the effect on ranging and command can be examined. Again the detectors are the most susceptible modules.

5.1.2.2 Ranging

The normal ranging output level from the detector is 16.7 millivolts rms. The detector dc line sensitivity is 67 mV/volt and 14 millivolts rms in the dc interference level.

$$\begin{aligned} \text{Interference level} &= -20 \log \left(\frac{16.7 \times 10^{-3}}{67 \times 10^{-3} \times 14 \times 10^{-6}} \right) \\ &= -85 \text{ dB} \end{aligned}$$

Similarly, for the command signal which is 12 millivolts rms worst case:

$$\begin{aligned} \text{Interference} &= -20 \log \left(\frac{12 \times 10^{-3}}{67 \times 10^{-3} \times 14 \times 10^{-6}} \right) \\ &= -82 \text{ dB} \end{aligned}$$

5.1.3 THRESHOLD PERFORMANCE AS A FUNCTION OF ERROR CHANNEL BANDWIDTH

The primary concern at threshold is boresight jitter which is a function of signal-to-noise ratio. The narrowest bandwidth is contained in the post-detection processing system.

Normalized jitter is

$$\phi \times S' = \frac{1}{\sqrt{2S/N(B_N)}}$$

where ϕ = Jitter

S' = System gain

(assume $S' = 0.5 \text{ (V/deg/V)}$)

Then

$$\begin{aligned} 20 \log \phi &= -20 \log 0.5 - 10 \log 2 - 10 \log (S/N) + 10 \log B_N \\ &= +6 - 3 - 10 \log (S/N) + 10 \log B_N \end{aligned}$$

At the receiver sum channel threshold (-155 dBm):

Noise = -174 dBm

Noise Figure = 5.5 dB

SNR = 13.5 dB/Hz

If the angle channels are 20 dB below the sum, and accounting for 3-dB TDM loss:

S/N in angle channel = -9.5 dB/Hz

Then $20 \log \phi = +9.5 + 10 \log B_N$

RMS jitter as a function of post-detection system bandwidth is shown in Figure 5-4.

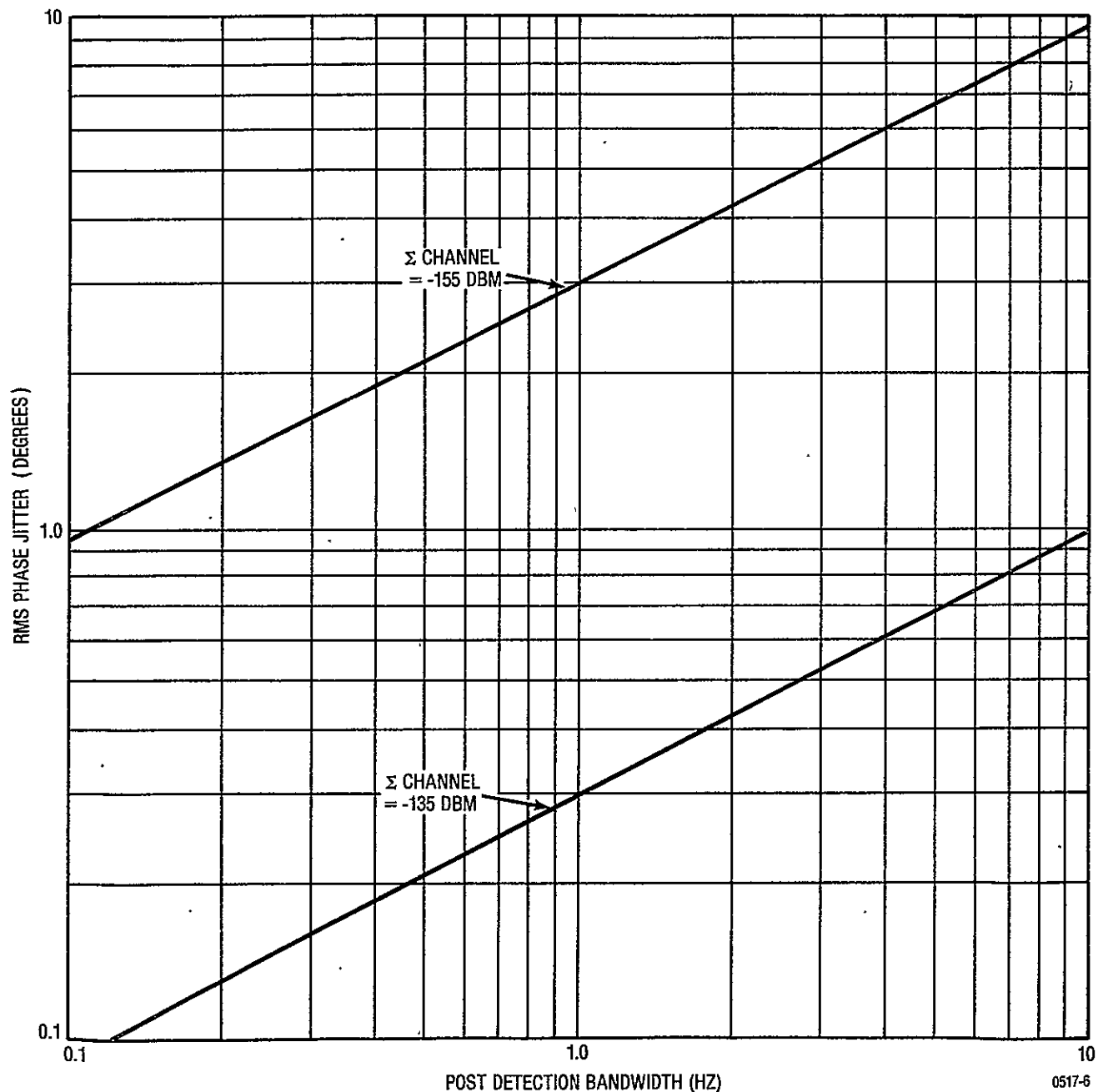


Figure 5-4. Antenna Phase Jitter vs. Post-Detection Bandwidth ($\Sigma/\Delta = 20 \text{ dB}$)

Figure 5-5 is a simplified block diagram of the Time Division Multiplexing system.



$e_s = A_s \cos \omega t$	Sum Channel	α_x	=	Precomparator X error
		α_y	=	Precomparator Y error
$e_x = A_x \cos (\omega t + \alpha_x)$	X Channel	α'_x	=	Postcomparator X error
		α'_y	=	Postcomparator Y error
$e_y = A_y \cos (\omega t + \alpha_y)$	Y Channel	K'_1 , etc	=	Postcomparator gain errors

Assuming no Doppler and no phase and amplitude errors,

$$K_1 = K_1', K_2 = K_2', \text{ and } \alpha_x' = \alpha_y' = 0$$

$$e_A = K_1 K_2 A_s \cos \omega_R t$$

$$e_B = K_1' K_2' A_x R_1(t) \cos(\omega_R t + \alpha_x) + K_1' K_2' A_y \bar{R}_1(t) \cos(\omega_R t + \alpha_y)$$

$$e_C = K_1' K_2' A_x R_1(t) R_2 \cos(\omega_R t + \alpha_x) + K_1' K_2' A_y \bar{R}_1(t) R_2 \cos(\omega_R t + \alpha_y)$$

$$e_D = [K_1' K_2' A_x R_1(t) R_2 \cos(\omega_R t + \alpha_x) + K_1' K_2' A_y \bar{R}_1(t) R_2 \cos(\omega_R t + \alpha_y)] K_3' [\cos(\omega_R t)]$$

$$= K_1' K_2' K_3' [R_1(t) R_2 A_x \cos \omega_R t \cos(\omega_R t + \alpha_x) + A_y \bar{R}_1(t) R_2 \cos(\omega_R t) \cos(\omega_R t + \alpha_y)]$$

if no ϕ error in channels

$$e_D = K_1' K_2' K_3' [R_1(t) R_2 A_y \cos^2 \omega_R t + A_y \bar{R}_1(t) R_2 \cos^2 \omega_R t]$$

$$= \frac{K_1' K_2' K_3'}{2} [R_1(t) R_2 A_x (\cos 2\omega_R t + 1) + \bar{R}_1(t) R_2 A_y (\cos 2\omega_R t + 1)]$$

- Filter $\cos 2\omega_R t$ term

$$e_D = \frac{K_1' K_2' K_3'}{2} [R_1(t) R_2 A_x + \bar{R}_1(t) R_2 A_y]$$

$$e_E = \frac{K_1' K_2' K_3'}{2} [R_1(t) R_2 R_3 A_x + \bar{R}_1(t) R_2 R_3 A_y]$$

$$\text{if } \bar{R}_1 = \frac{1}{R_1}$$

$$R_2 = R_3 = R_1(t)$$

Where $R_1(t) = R_1$ delayed through the system

then

$$e_E = \frac{K_1' K_2' K_3'}{2} R_1(t) R_2 R_3 A_x = \frac{K_1' K_2' K_3'}{2} A_x$$

and

$$e_F = \frac{K_1' K_2' K_3' K_4'}{2} A_x$$

Let α_o = Sum channel tracking error with Doppler

$\alpha_1, \alpha_2, \alpha_3$ = Sum channel phase shifts

$\alpha_1', \alpha_2', \alpha_3'$ = Angle channel phase shifts

α_x' = Differential phase shift sum to X channel

$K_1 K_2 K_3$ = Sum channel gains

$K_1'K_2'K_3' = \angle$ channel gains

$K' =$ Differential gain sum to $\angle$ channel

Then:

At E

$$e_E = K_1'K_2'K_3' [R_1(t)R_2R_3A_x \cos (\omega_r t + \alpha_x') \cos (\omega_r t + \alpha_o) \\ + \overline{R_1}(t)R_2R_3A_Y \cos (\omega_r t + \alpha_y') \cos (\omega_r t + \alpha_o)]$$

$$e_E = \frac{K_1'K_2'K_3'}{2} \left\{ R_1(t)R_2R_3A_x [\cos (2\omega_r t + \alpha_x' + \alpha_o) + \cos (\alpha_x' - \alpha_o)] \right. \\ \left. + A_Y \overline{R_1}(t)R_2R_3 [\cos (2\omega_r t + \alpha_y' + \alpha_o) + \cos (\alpha_y' - \alpha_o)] \right\}$$

- Filter $\cos 2\omega_r t$ term

$$e_E = \frac{K_1'K_2'K_3'}{2} R_1(t)R_2R_3A_x \cos (\alpha_x' - \alpha_o) + \overline{R_1}(t)R_2R_3A_Y \cos (\alpha_y' - \alpha_o)$$

Then errors are:

1. X-Y Crosstalk = $\overline{R_1}(t)R_2R_3A_Y \cos (\alpha_y' - \alpha_o)$ term, which is a function of R_1 and $\overline{R_1}$ overlap (rise/fall time and symmetry and the attenuation of Switches S_1 , S_2 , and S_3 . Also the delay differences between the switch function R_1 as it is delayed in the IF path and R_2 and R_3 .
2. Gain tracking = $K' = K_1'K_2'K_3' - K_1K_2K_3$
3. Phase tracking = $\cos (\alpha_x' - \alpha_o)$

where $\alpha_x' = \alpha_x - \alpha_{xE}$

where $\alpha_x =$ true angle, $\alpha_{xE} =$ measured angle

The X-Y crosstalk isolation is specified to be greater than 30 dB. Crosstalk is created by:

- Absolute attenuation of the undesired signal in the multiplexer
- Overlap of rise and fall times of the multiplexer and sample and hold circuitry
- Difference in delay time of the multiplexed signal and the reference chopping frequency of the sample and hold circuits
- Symmetry of all waveforms

Examination of these contributions yields the following:

- Absolute Attenuation

Switch S_1 provides the ON/OFF attenuation ratio and is specified to provide at least 40 dB of attenuation to OFF signal.

- Rise and Fall Time Overlap

The rise and fall times of all switches and sampling signals are specified to be less than 100 nanoseconds. The maximum amount of crosstalk due to signals overlapping during these time periods is:

$$10 \log \frac{200 \times 10^{-9}}{10 \times 10^{-3}} = 47 \text{ dB}$$

- Delay Time

The 50 Hz IF chopping frequency is stretched by the rise and fall response of the IF filters. The primary contribution is made by the prediction filter which has a 3 kHz 3 dB response.

$\tau_r = 53 \mu_s$ for this filter.

If all sample pulses are coincident, the crosstalk level will be the energy contained in the fall time characteristics of the undesired signal as shown in Figure 5-6.

The rms voltage of this signal over the sample period is equal to

$$A_{rms} = \left[\frac{A_{pk}^2}{T} \int_0^T e^{-2T/\tau} dt \right]^{1/2}$$

For $\tau = 53 \times 10^{-6} \text{ s}$ and a sample time of $10 \times 10^{-3} \text{ s}$ ($T = 189$) then $A_{rms} = 0.0026 A_{pk}$
and Crosstalk = $20 \log 0.0026 = 51.5 \text{ dB}$

- Waveform Symmetry

The difference in duty cycle between R and R is less than 5 nanoseconds. This crosstalk contribution, then, is greater than 60 dB.

- Summary

The limiting contributor to crosstalk is the S-Band switch which provides a minimum of 40 dB of isolation.

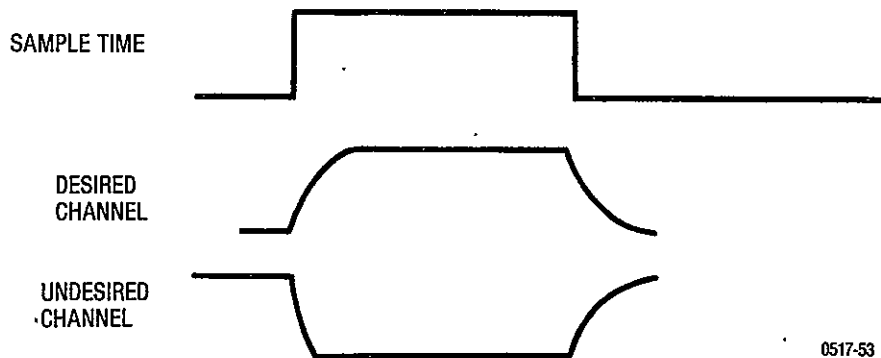


Figure 5-6. Signal Characteristics

5.1.4.1 Error Channel Gain Tracking

The error channel must gain track the sum channel over the signal dynamic range and over temperature. The primary contributor to tracking with dynamic level change is the first IF which contains all AGC amplifiers. All amplifiers are selected for gain versus AGC voltage. In addition, separate AGC voltage level and slope controls are provided to ensure optimum tracking (± 0.1 dB goal).

All of the following modules contribute to temperature performance.

They are:

- Preselector
- Mixer preamplifier
- First IF
- Second IF
- Detector

Tracking accuracy must be within ± 0.2 dB over the temperature range -10°C to $+55^{\circ}\text{C}$ for a constant signal level. The following data was taken from standard and microminiature test results.

- Preselector

Gain: -1 dB

Δ Temperature: ± 0.1 dB

Tracking contribution: ± 0.1 dB

- Mixer Preamplifier

Δ Gain: +25 dB

Δ Temperature: ± 1 dB

Tracking: ± 0.8 dB

- First IF

Gain: +52, -49

Δ Temperature: +2, -12 dB

Tracking: ± 2 dB

- Second IF

Gain: +31 dB

Δ Temperature: ± 3 dB

Tracking: ± 0.5 dB

- Detector .

Gain: +15 dB

Δ Temperature: ± 0.5 dB

Tracking: ± 0.5 dB

Total: ± 4.0 dB tracking error peak

2.25 dB tracking error rms

The above data indicates that temperature compensation is required to meet ± 0.2 dB gain tracking. This will be accomplished by temperature compensating the gain control part of an IF amplifier stage.

5.1.4.2 Error Channel Phase Shift

The major contributions to phase shift in the error channels are attributed primarily to the filtering networks. Temperature data from the Microminiature and Standard Transponder module data was used for determination of expected phase shift.

- Preselector: 5 pole; 60 MHz, 3 dB bandwidth; 0.1 dB ripple

Test results: Δf_o with temperature = 3.5 MHz from -20°C to $+75^\circ\text{C}$

$$\theta = \frac{\Delta f}{B_{3dB}} \times 400 \text{ for 5 pole 0.1 dB filter*}$$

$$\theta = \frac{3.5}{60} \times 400 = 23^\circ$$

The filters are expected to track within 10 percent, therefore $\Delta\theta = 2.3^\circ$.

- Mixer-preamplifier: 3 pole; 8 MHz, 0.1 dB ripple filter*

Test results: $\Delta f_o = 0.4$ MHz from -20°C to $+75^\circ\text{C}$

$$\theta = \frac{\Delta f}{B_{3dB}} \times 180 \text{ for 3 pole 0.1 dB filter}$$

$$\theta = \frac{0.4}{8} \times 180 = 9.0^\circ$$

The filters are expected to track within 20 percent, therefore $\Delta\theta = 1.8^\circ$.

- First IF: 15 MHz low pass filter

Test results: $\Delta f = 0.25$ MHz for -20°C to $+75^\circ\text{C}$

Expected Δf tracking = 10 percent

$\therefore \Delta f = 0.025$ MHz

$$\theta = \tan^{-1} 2 \left(\frac{0.025}{15} \right) = 0.2^\circ$$

*Motorola Technical Memo RFS-55, S. G. Miller

- Second IF: 3 kHz single pole filter test result percent, Δf tracking between units = ± 50 Hz for -20° to $+75^\circ\text{C}$

$$\theta = \tan^{-1} 2 \left(\frac{\omega - \omega_0}{\omega_3} \right) = \tan^{-1} 2 \left(\frac{100}{3000} \right) = 3.8^\circ$$

- 250 kHz command filter

Test results: $\Delta f = 20$ kHz

Assume tracking of 20 percent, $\Delta f = 4$ kHz, $\Delta\phi = \tan^{-1} 2 \times (4/250) = 1.8^\circ$

Total estimated phase shift = 9.9°

Total estimated rms phase shift = 5.1°

5.1.5 ANGLE CHANNEL LINEARITY

The angle channel gain distribution is shown in Figures 5-7 and 5-8. The system is designed to accommodate three sigma noise peaks linearly throughout the channel.

The output peak noise S-curve is 5 volts and the signal S-curve is 100 millivolts peak for an angle channel input 10 dB below the sum channel.

5.1.6 CHOPPING FREQUENCY ANALYSIS

The chopping frequency selection is constrained by two parameters. The first constraint is the final IF bandwidth. This 3-dB bandwidth is 3 kHz. In order to preserve the square wave, the fifth harmonic as a minimum must be passed through the filter. There, $f_{\max} = 1500/5 = 300$ Hz.

The second constraint is the allowable amount of chopping signal allowed at the output. Applying the criteria that the output shall be no greater than 2 percent of the level that occurs when the angle channel input signal is equal to the sum channel input, then this value is 0.02×316 millivolts or 6.32 millivolts. The maximum input to the sample and hold circuit will be 44 millivolts when the sum channel is equal to the error channel. The required attenuation will be $20 \log (6.32 \text{ mV}/44 \text{ mV}) = 17$ dB.

Assuming a maximum post-detection noise bandwidth ($B_n = 20$ Hz), the 3-dB frequency will be 6.35 Hz. To realize approximately 17 dB of attenuation, the chopping frequency should be at least 3 octaves higher than 6.35 Hz, therefore, a lower bound of 50 Hz is set.

The bounds on chopping frequency are thus $50 \leq f_{\max} \leq 300$ Hz. From an X-Y channel crosstalk standpoint the minimum frequency (maximum period) is desirable since this minimizes the effects of time delay, rise times, etc. Therefore, 50 Hz will be chosen as the chopping frequency.

5.2 Quadrature

Figure 5-9 is a simplified block diagram of the quadrature phase multiplexed system, from which the error channel crosstalk, linearity, drift, and behavior under signal dynamics and temperature will be analyzed.

$$e_s = A_s \cos \omega_i t$$

$$e_x = A_x \cos (\omega_i t + \alpha_x)$$

$$e_y = A_y \cos (\omega_i t + \alpha_y)$$

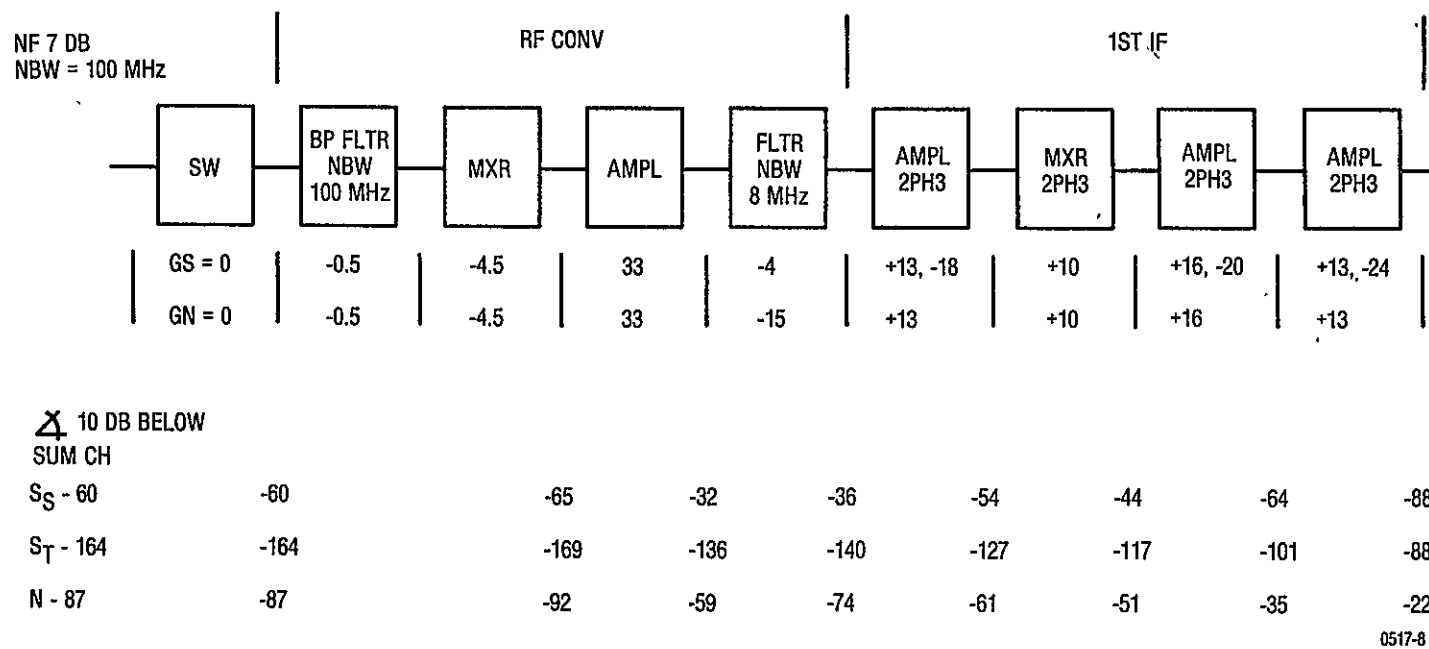


Figure 5-7. TDM Angle Channel Gain Distribution (RF Converter and First IF)

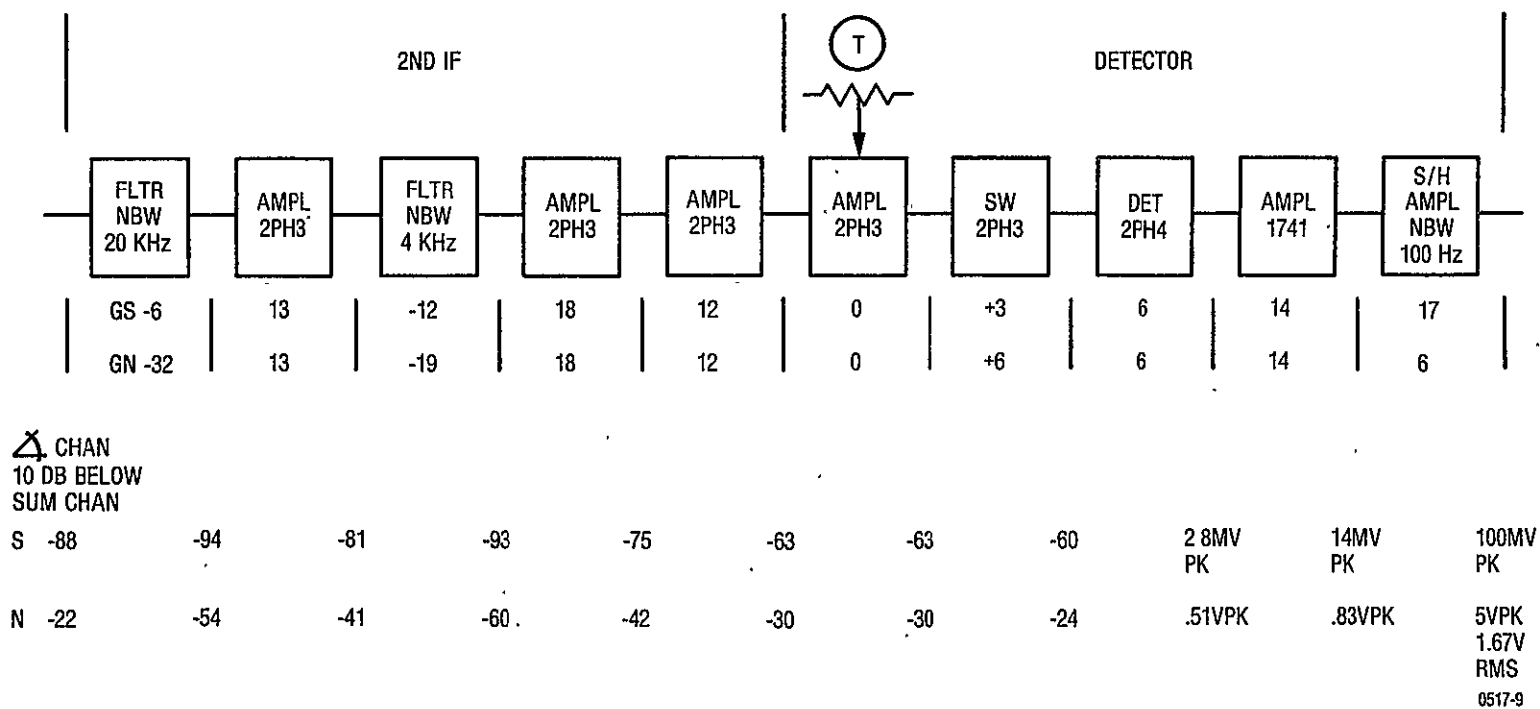


Figure 5-8. TDM Angle Channel Gain Distribution (Second IF and Detector)

α_u = Phase shift from 90° in summer

α_d = Doppler phase shift

α_n' = Channel phase shifts

K_n' = Channel gain

$$e_A = K_1 K_2 A_s \cos \omega_r t$$

$$e_B = K_1' [A_x \cos (\omega_r t + \alpha_x) + A_y \cos (\omega_r t + \alpha_y + \pi/2 + \alpha_u)]$$

$$= K_1' [A_x \cos (\omega_r t + \alpha_x) + A_y \sin (\omega_r t + \alpha_y + \alpha_u)]$$

$$e_C = K_1' [A_x \cos (\omega_r t + \alpha_x) + A_y \sin (\omega_r t + \alpha_y + \alpha_u)]$$

$$e_D = K_1' K_2' K_3' K_4' \cos \omega_r t [A_x \cos (\omega_r t + \alpha_x) + A_y \sin (\omega_r t + \alpha_y + \alpha_u)]$$

$$= \frac{K_1' K_2' K_3' K_4'}{2} \{ A_x [\cos (2\omega_r t + \alpha_x) + \cos \alpha_x]$$

$$+ A_y [\sin (2\omega_r t + \alpha_y + \alpha_u) + \sin (\alpha_y + \alpha_u)] \}$$

5.2.1 FILTER $2\omega_r t$ TERMS

$$e_D = \frac{K_1' K_2' K_3' K_4'}{2} [A_x \cos \alpha_x + A_y \sin (\alpha_y + \alpha_u)]$$

When Doppler phase shift (α_d) and sum-to-angle channel phase tracking errors ($\Delta\alpha$) are considered;

$$e_D = \frac{K_1' K_2' K_3' K_4'}{2} [A_x \cos (\alpha_x + \alpha_d + \Delta\alpha) + A_y \sin (\alpha_y + \alpha_u + \alpha_d + \Delta\alpha)]$$

Then $\Delta K = K_1 K_2 K_3 K_1' K_2' K_3' K_4'$ is the gain tracking error.

$\alpha_D + \Delta\alpha$ are phase tracking errors and

$A_y \sin (\alpha_y + \alpha_u + \alpha_d + \Delta\alpha)$ is the crosstalk term.

- Error Channel Gain Tracking

The gain tracking analysis of the TDM system applies here as well. Temperature and AGC compensation are required to constrain the maximum variation to ± 0.3 dB.

- Error Channel Phase Tracking

The phase tracking analysis of the TDM system applies in this case also. A 5-degree RSS change is expected with temperature variations with an additional 5 degrees due to Doppler.

- Angle Channel Crosstalk

In the Quadrature System phase shifts also affect channel crosstalk. The phase contributions are due to sum-to-error channel tracking differences over temperature, phase shift with Doppler and nonorthogonality of the X and Y channels. Assuming a ± 10 degree phase shift, the error channel output is

$$A_p (\cos 10 \pm \sin 10)$$

$$E = (0.985 \pm 0.17; A_p = 1.159 A_p \text{ or } 0.811 A_p)$$

$$\text{Crosstalk} = 20 \log \frac{0.174}{0.985} = -15 \text{ dB}$$

A 30-dB crosstalk requirement equates to a total drift and nonorthogonality of 1.8 degrees.

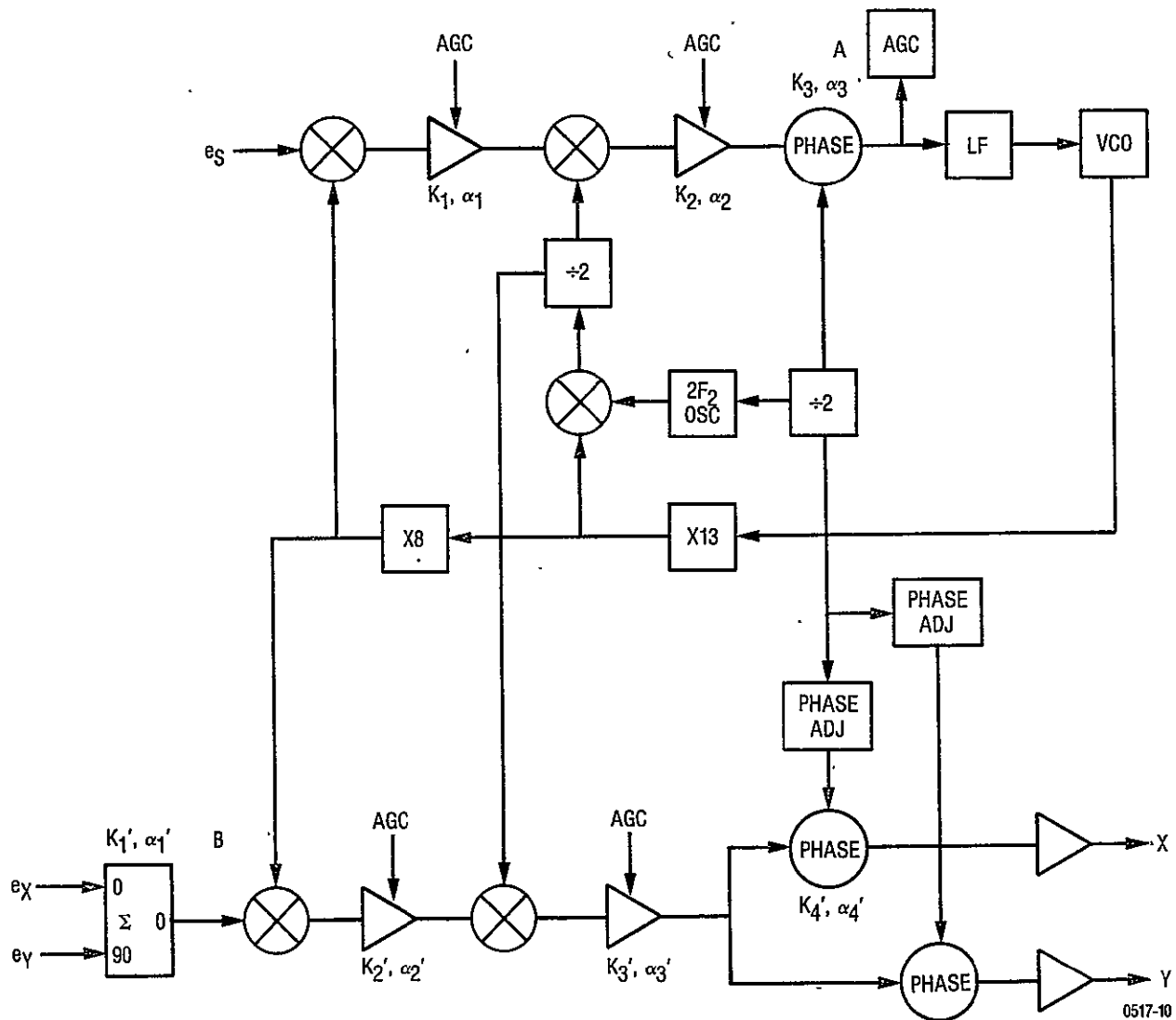


Figure 5-9. Quadrature Receiver Functional Block Diagram

SECTION 6

6. MODULE DESCRIPTION

6.1 S-Band RF Switch

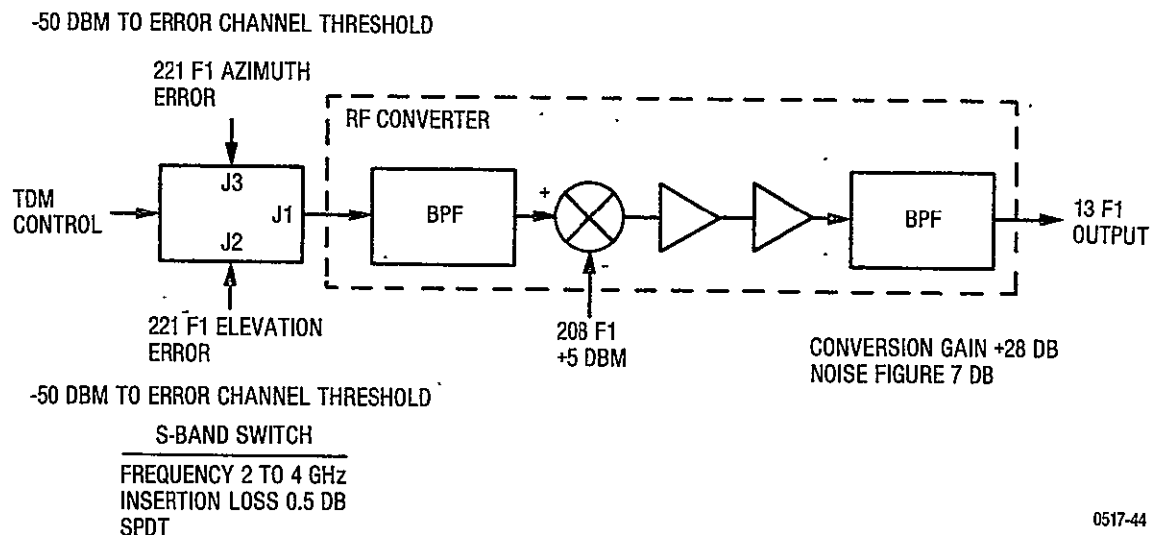
The S-Band RF switch is a General Microwave, Series F8922, option 27 single-pole double-throw switch. It has a broadband frequency range of 2 to 4 GHz with a maximum insertion loss of 0.5 dB, and 60 dB of isolation between ports. The logic control input is TTL compatible and receives a 50 percent duty cycle clock pulse from the synchronous demodulator/control module at a 50 Hz sampling rate. The switch loss will add 0.5 dB to the system noise figure.

6.2 RF Converter

The RF converter is common to both the TDM System and quadrature system. Figures 6-1 and 6-2 show block diagrams of the two systems. The only difference between the two is that S-Band RF switch (TDM System) is replaced with a 90-degree hybrid (Quadrature System). The RF converter consists of three modules: preselector, balanced mixer, and IF preamplifier. The preselector provides filtering of the input signal and prevents the generation of spurious frequencies in the mixer. The preselector also provides reverse isolation at the LO frequency.

The single balanced mixer consists of a microstrip ring configuration such that the LO signal path is one-half wave length longer to one diode than to the other diode. The RF signal is inphase at the diodes which results in a summation of RF power and cancellation of the LO noise. The IF preamplifier consists of a two-stage transistor amplifier followed by a three-pole, capacitor-coupled bandpass filter.

Conversion gain of the RF converter is +28 dB with a 7-dB noise figure. The RF converter is identical to the RF converter used in the sum channel to minimize gain and phase variations over temperature and input signal dynamic range.



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Figure 6-1. TDM S-Band Sampling Switch and RF Converter Block Diagram

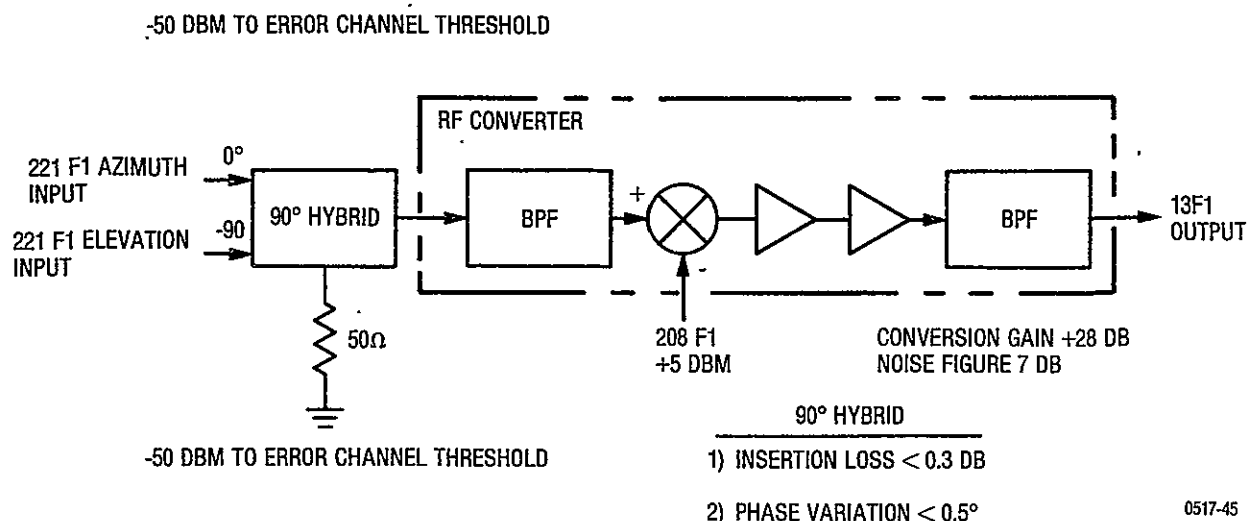


Figure 6-2. Quadrature RF Converter Block Diagram

6.3 First IF

The first IF module amplifies the $13F_1$ signal from the IF preamplifier and mixes this signal with the second LO signal, $13F_1 - F_2$, to yield an F_2 signal. This F_2 signal is then passed through an 18.3-MHz low pass filter and amplified (see Figure 6-3).

AGC voltage input from the sum channel transponder is modified for gain-slope characteristics before being impressed on the 76T amplifiers. R21 and R28 adjust the gain and slope of the AGC voltage with an active 1747 dual operation amplifier which also provides AGC isolation between the sum channel and the error channel. This part of the AGC circuit design may be implemented with the use of one operational amplifier instead of two that were used for the breadboard. However, there is a definite necessity for at least one amplifier in the angle channel AGC circuitry, for the gain and slope calibration of the angle channel AGC voltage, so that the system output gain variations comply with the 5 percent specification. Module AGC dynamic range is 110 dB (± 55 dB), with an AGC amplifier sensitivity of approximately 100 dB per volt.

The filtered F_2 (12.250 MHz) output is held constant at -88 dBm. The LO 2 ($13F_1 - F_2$) signal is derived from the sum channel transponder at a power level of -21 dBm which is amplified before mixing.

The IF section of the first IF module is identical to the NST first IF module with close gain matching of the 2PH1 and 2PH3 devices. All other devices used (2PH3 amplifier and operational amplifiers) are the same as existing NST device usage.

6.4 Second IF (Figure 6.4)

The second IF module contains two crystal bandpass filters centered at the F_2 frequency, with a 20-kHz and 4.7-kHz noise bandwidth, respectively. The F_2 input signal is derived from the output of the First IF module. The F_2 signal is separated after filtering into the azimuth amplifier error channel and the elevation amplifier error channel. Both outputs are balanced and are fed to the detector at a power level of -63 dBm. For this breadboard version, two

channels were used to aid in gain definition through separation of channels. However, it is considered to be redundant and one channel will suffice. Further detailed explanation is covered in the detector module description (paragraph 6.5). Filtering in the 3-kHz bandwidth filter will cause the 50-Hz sampled RF waveform to be degraded by the time constant of the filter phase slope. One time constant is 53 microseconds, consequently the RF pulse train will be degraded by 212-microsecond rise and fall times.

Module gain is +25 dB with close matching of the 2PH3 amplifiers to that of the NST second IF module for gain tracking.

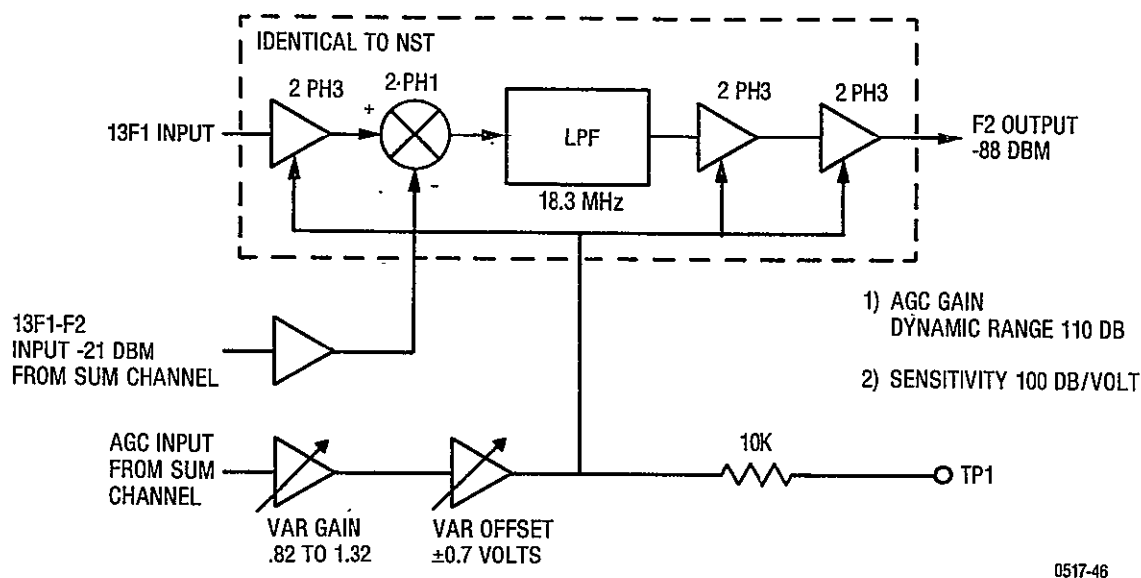


Figure 6-3. First IF Module Block Diagram

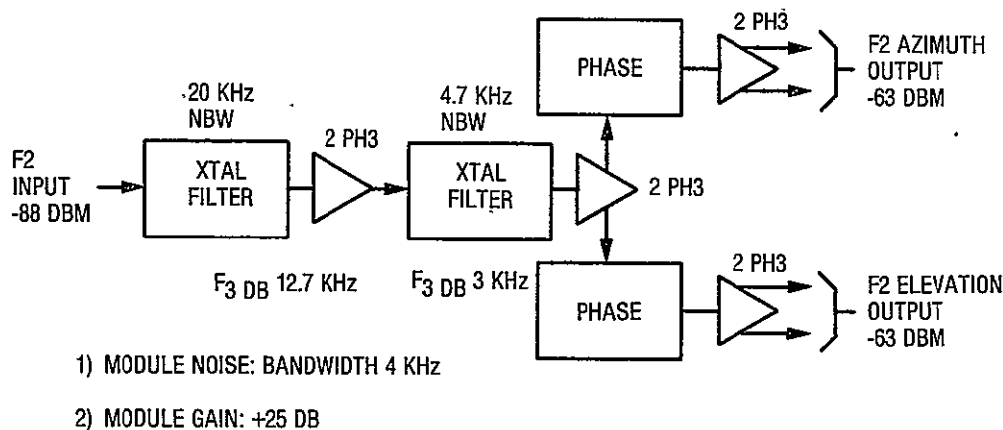


Figure 6-4. Second IF Module Block Diagram

6.5 Detector

Two types of detection systems have been employed for the Monopulse Receiver Study, Time Division Multiplexing (TDM) and Quadrature. However, each type of detection system requires two separate modules, with an F_2 reference amplifier module being common to both types of detectors with built in phase adjustments to accommodate phasing requirements of both detection systems. Both types of detectors are similar in the use of 2PH3 and 2PH1 devices as used in the NST detector module. For the breadboard version, discrete non-beam-lead counterparts were used (76T RF amplifiers and X336R phase detectors).

6.5.1 F_2 REFERENCE AMPLIFIER (FIGURE 6-5)

The F_2 reference amplifier receives its signal from the sum channel transponder at a power level of -48 dBm. This signal is amplified to -8 dBm and split into the azimuth and elevation reference channels. In each reference channel, two phasing adjustments, each with a range of ± 55 degrees, have been included. For the azimuth channel, T1 and T2 have a combined phase adjustment range of ± 110 degrees, also T3 and T4 for the elevation channel. Each phasing network has a 50-ohm characteristic impedance at center frequency with a phase angle of 0 degree. Phasing adjustments may be made from outside of the module with the use of potentiometers. The reference signals are then amplified and fed to a lattice network for 0, 90, or 180 degree coarse phase adjustments. Output power is approximately -5 dBm, which is fed through balanced lines to the detector module reference ports. The lattice phase network was included for additional latitude in phasing requirements. However, it is now felt that these circuits are redundant and may be omitted due to the ± 110 -degree range of the variable phase networks.

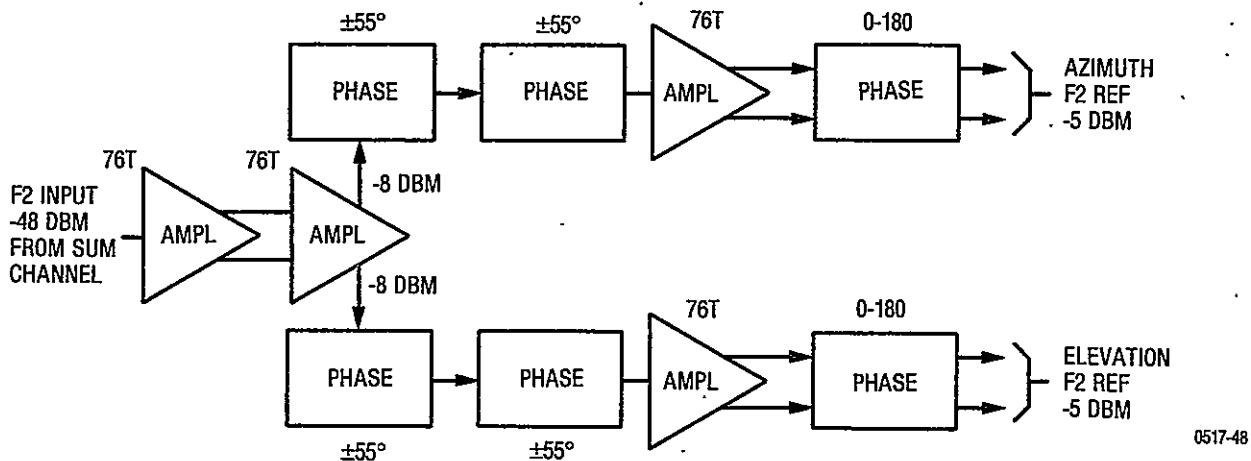


Figure 6-5. F_2 Reference Module Block Diagram

6.5.2 TDM DETECTOR (FIGURE 6-6)

The TDM detector employs two balanced sinusoidal coherent amplitude detectors (CADs) with a detector sensitivity, K_D , of 2.5 volts per radian. Input signal power from the second IF module of -63 dBm is amplified and separated into the azimuth and elevation signals prior to detection. As described previously in the second IF module description, the breadboard version separates the F_2 IF's into two channels prior to the detector module. Each channel, however, still contains both azimuth and elevation error information with real pre-detection separation of

these signals occurring in U1 and U2 chopping amplifiers. It was felt that the separation of the IF signal into two channels is not necessary prior to pre-detection chopping. The effect will reduce circuitry in the second IF module and at the input to the detector module with just a single temperature gain compensating amplifier common to both the azimuth and elevation error channels. However, all circuits that follow the temperature gain compensation amplifier will be required to match in gain, and track correctly to ensure the azimuth and elevation system outputs be contained within the 5 percent specification.

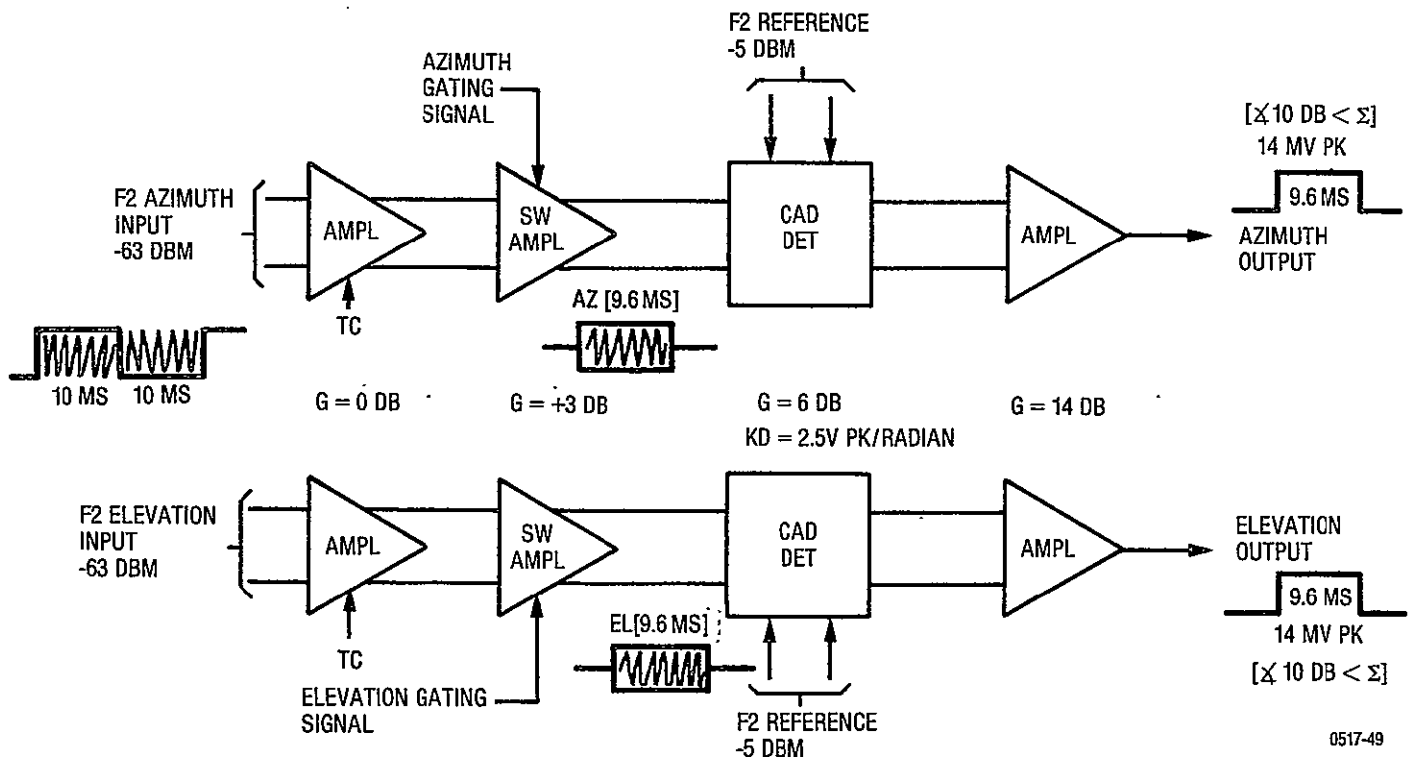


Figure 6-6. TDM Coherent Amplitude Detector Module Block Diagram

The need for two detectors was determined to be necessary because of the system output requirement of 2 percent dc offset balance. The effect of the dc chopper stabilized detectors prevents the detector system dc offsets from contributing to the system output balance. Without the chopper stabilized detection system, the output would be effected as follows: Given a well-balanced 336 detector with 50 dB of balance, its offset contribution would be ± 1.6 millivolts. This offset is then amplified in the 1741 operational amplifier (assume operational amplifier to be ideal $V_{os} = 0$) to $5(\pm 1.6 \text{ millivolts}) = 8$ millivolts. This offset then would be processed through the sample and hold module to the system output with an additional gain of 7 (16.9 dB), contributing 56 millivolts to the output dc offset, (assuming the output LM108 operational amplifier to be ideal $V_{os} = 0$). With a maximum system boresight error output voltage of 316 millivolts, of which 2 percent = 6.3 millivolts, 56 millivolts would degrade this specification to 18 percent. Hence the need for pre-detection chopping with two detectors is a real requirement. The predetection signal gain from the input to the module to the detector input is +6 dB, with a 3 dB loss due to signal separation. Therefore, total signal gain prior to detection is 3 dB or -60 dBm of delivered signal power to the detector input (0.7 millivolt rms on each side of a balanced input line). U11 and U12 perform gain compensation due to temperature with the aid of a simple

resistive/thermistor positive TC network. U1 and U2 are the azimuth and elevation chopping amplifiers which are switched On/Off with the correctly phased switching waveform of 0 volt to -4 volts. On/Off isolation is approximately 60 dB.

Input noise power at threshold is -30 dBm which is enhanced by a +6 dB gain to -24 dBm, (45 millivolts rms or 135 millivolts peak each side of a balanced line); prior to detection. Post-detection noise power is 0.83 volt rms or 2.5 volts peak. Both balanced detectors are loaded into differential operational amplifiers with a gain of five and low pass filtering of 33.6 kHz. The outputs of the detectors are ac-coupled into the TDM sample and hold/control module.

6.5.3 QUADRAPHASE DETECTOR

The Quadrature Detection System employs two balanced sinusoidal coherent amplitude phase detectors, with both the azimuth and elevation F_2 references phased orthogonally. The S-Band signal condition requirements necessitate the use of a 90-degree hybrid for generation of the azimuth and elevation orthogonal vectors.

F_2 signal input power and signal gain factors for pre-detection are identical to the TDM System. However, unlike the TDM System which chops each channel in sequential time slots, the Quadrature Detection System chops both channels simultaneously. The chopping of the IF signal and de-chopping of the error voltages essentially provide a chopper-stabilized detection system.

The azimuth and elevation error outputs are ac-coupled into a synchronous demodulator with signal processing identical to the TDM System (see paragraph 6.6.1). System output is identical to the TDM System outputs which are related to the S-Band null depth for pointing accuracy.

6.6 Sample and Hold/Timing Control Module

6.6.1 TIMING

Both types of systems utilize the same Motorola CMOS low power, high noise immunity devices for logic timing and generation of control signals. The CMOS digital family was chosen for the above characteristics and because of implementing the design into a radiation hardened RCA metal gate MOS circuits. Due to beamlead limitations for the CMOS family, use of RCA flatpacks would be a viable alternative. This report does not preclude the possibility of other digital families such as TTL or DTL, however, there are design tradeoffs to be considered such as power consumption and noise immunities. Figure 6-7 and Figure 6-8 are the block diagrams for the two systems.

6.6.2 TIME DIVISION MULTIPLEXING

6.6.2.1 Digital Control Signal Generation

The master oscillator is a programmable, free-running RC oscillator set at 51.2 kHz by R5. The clock frequency is internally divided by 2^{10} (1024) to a baseband frequency of 50 Hz for both frequency stability and symmetry of the output squarewave. For this application the stability of the oscillator will be as good as the RC temperature coefficients, however, high stability is not really a requirement.

Given R	= 110 k Ω ohms $\pm$ 50 ppm/ $^{\circ}$ C
C	= 470 pF NPO $\pm$ 30 ppm/ $^{\circ}$ C
ΔT	= $\pm$ 35 $^{\circ}$ C
Total RC variation	= $\pm$ 0.28%

Frequency variation

$$= 51.2 \text{ kHz (0.0028)}$$

$$= \pm 143 \text{ Hz}$$

Baseband frequency variation

$$= \frac{\pm 143}{1024} = \pm 0.14 \text{ Hz}$$

Given worst case conditions, the baseband period will vary from 20.056 milliseconds, ($T/2 = 10.028$ milliseconds) to 19.944 milliseconds ($T/2 = 9.972$ milliseconds). Therefore, a maximum worse case specification of ± 28 microseconds will place a lower boundary as the minimum guard time of the sampling pulse.

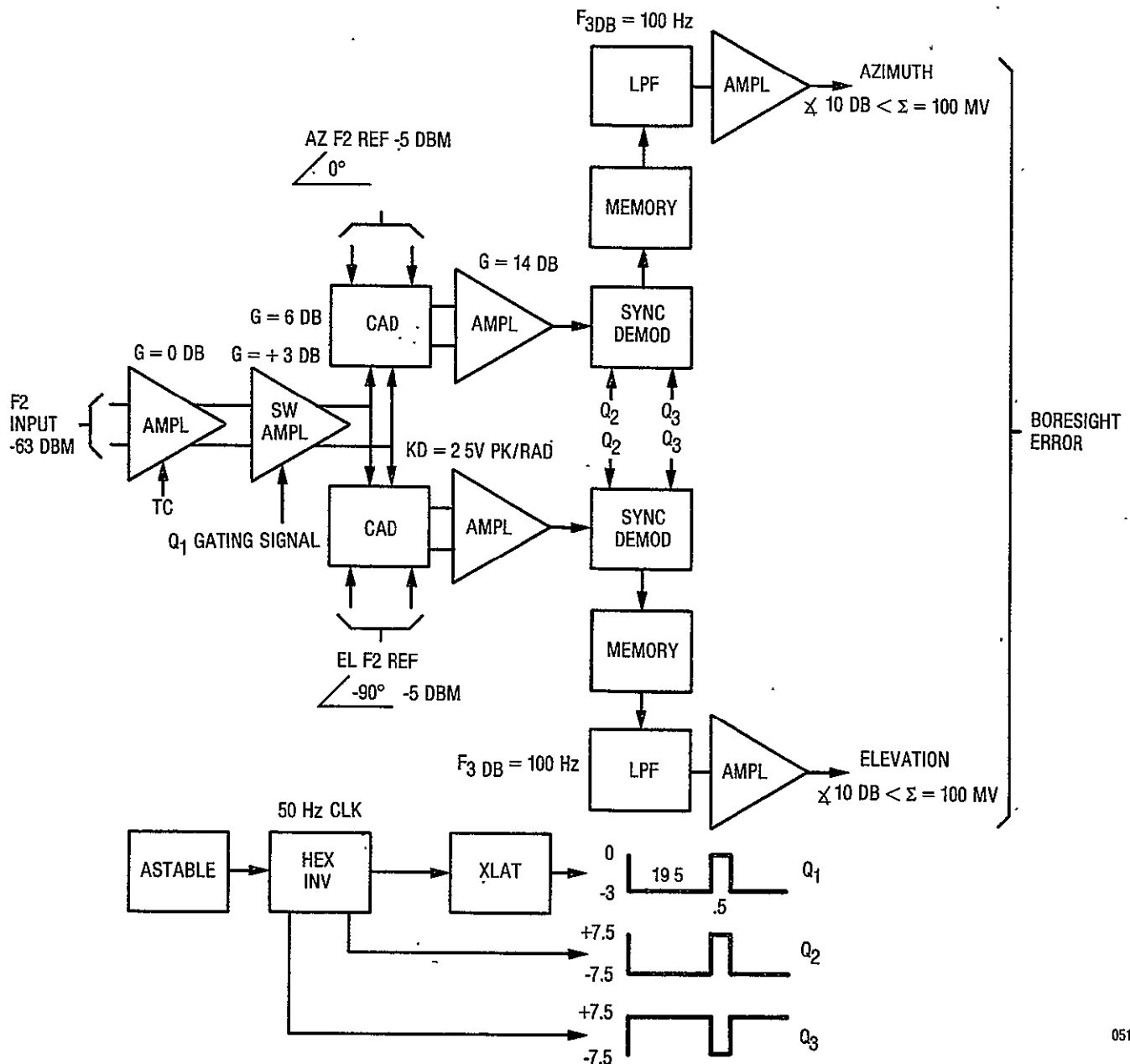


Figure 6-7. Quadrphase Detection/Timing and Sampling System Module Block Diagram

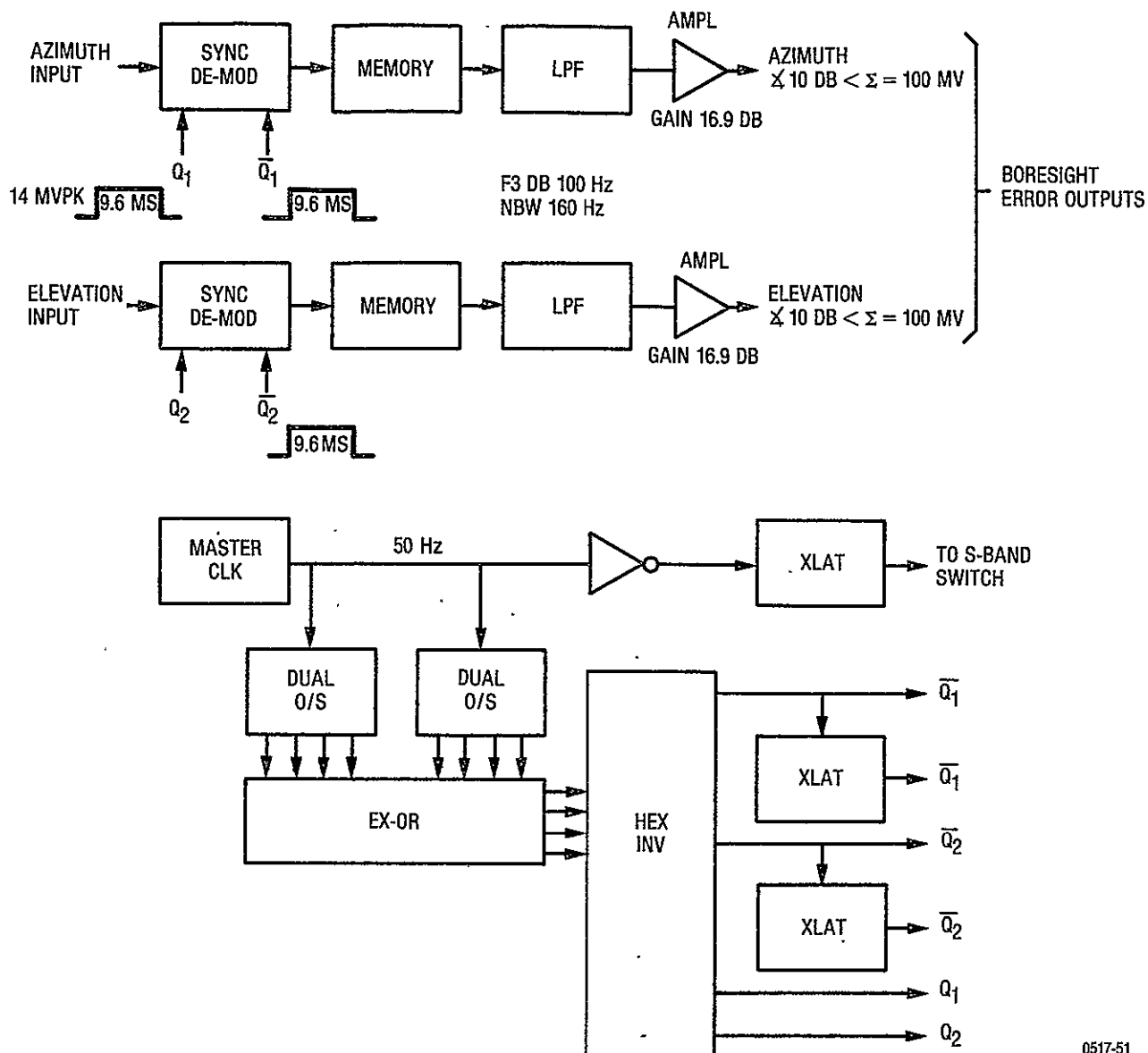


Figure 6-8. TDM Sample and Hold/Control Module Block Diagram

The clock pulse is then fanned out to two dual retriggerable monostable multivibrators with U2 clocked on the leading edge and U3 clocked on the falling edge. Each monostable provides a 200-microsecond and a 9.8-millisecond blanking pulse with U2 operating in the first half-period and U3 in the second half-period. The one-shot outputs provide the blanking pulses for gating in the exclusive-OR gates which are then inverted and buffered to external circuitry. Both the master clock and one-shots are preset to a logic 0 (low-level) with power turn-on. Switching waveforms Q_1 , $\overline{Q_1}$, Q_2 , and $\overline{Q_2}$ are of the correct phasing for chopping of the azimuth and elevation channels with a guard time of 200 microseconds on each side of the sampling period. The 200 microseconds guard time on the rising edge and falling edge of the signal removes the rise and fall time degradation of the signal as previously discussed in the second IF narrowband filter and minimizes crosstalk between channels with less than 0.177 dB loss of signal power.

Network synthesis for the generation of guard times was implemented with RC timing for design reasons of both simplicity of circuits with a minimal number of digital packages and for latitude in guard time variation. An alternative approach of digitally synthesizing the required 9.6-millisecond pulse with 200-microsecond guard times may be implemented digitally with a higher clock frequency and a more complex implementation of dividers and gates.

The limitations of the 9.8-millisecond blanking pulse width are 9.972 milliseconds due to frequency stability requirements. The leading edge of the 9.6-millisecond sampling pulse will have an insignificant change over temperature. Consequently, the falling edge of the 9.6-millisecond sampling pulse is of concern and will be constrained to a minimum variation due to temperature with the use of low temperature co-efficient resistors and capacitors that form the RC timing networks.

Given	R	=	280 k Ω $\pm$ 50 ppm/ $^{\circ}$ C
	C	=	0.1 μ F NPO $\pm$ 30 ppm/ $^{\circ}$ C
	Δ T	=	$\pm$ 35 $^{\circ}$ C

Variation in R	=	280 k Ω $\pm$ 490 Ω
	=	280 k Ω $\pm$ 0.175%

Variation in C	=	0.1 μ F $\pm$ 0.105%
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Total pulse width variation	=	9.8 ms $\pm$ 0.28%
-----------------------------	---	--------------------

Worst case pulse width variation	=	9.8 ms $\pm$ 20 μ s
----------------------------------	---	-------------------------

- Total Sampling Pulse Width Variation

Pulse width (nominal)	=	9.8 ms - 0.2 ms - 9.6 ms
-----------------------	---	--------------------------

Pulse width (Maximum)	=	(9.8 + 0.028) - 0.2 = 9.628 ms
-----------------------	---	--------------------------------

Pulse width (Minimum)	=	(9.8 - 0.028) - 0.2 = 9.572 ms
-----------------------	---	--------------------------------

- Signal Loss In Power

$$10 \log \frac{PW}{10 \text{ ms}}$$

Pulse width (nominal)	=	10 log 0.960	=	-0.177 dB
-----------------------	---	--------------	---	-----------

Pulse width (maximum)	=	10 log 0.9572	=	-0.190 dB
-----------------------	---	---------------	---	-----------

Pulse width (minimum)	=	10 log 0.9628	=	-0.165 dB
-----------------------	---	---------------	---	-----------

Total variation in system output power due to temperature is $\pm$ 0.012 dB. Optimal design considerations required the CMOS logic circuitry to be operating from $\pm$ 7.5 volts dc power. These supply voltages are zener regulated from the available $\pm$ 12 volt dc line. Minimization of the ON resistance of the MC14066 quad analog transmission gates required a control voltage pulse of $\pm$ 7.5 volts dc. ON resistance is typically 75 ohms with least temperature variations. All other pulse level requirements have been synthesized using a differential "totem-pole" output translator. The differential mode design was selected because of its constant current source characteristics, thereby minimizing current spikes on the power lines with the fast transition times of the clock pulse. The totem-pole output provides fast rise and fall times with active pull-up and pull-down characteristics. The S-Band modulating switch requires a TTL-compatible pulse of logic 0 (0 volt dc) to a logic 1 (+5.0 volts dc).

The two 76T chopping amplifiers in the detector module prior to detection requires pulses of 0 volt (amplifier Off) to -4 volts (amplifier On). Three translators accomplish the generation of these signals with all control signals being in synchronism with the same source and being referenced to the S-Band modulating switch control. Potentiometers R1 and R3 adjust the guard times for Q_2 and $\bar{Q}_2$, while R2 and R4 similarly adjust the Q_1 and $\bar{Q}_1$ waveform guard times.

6.6.2.2 Synchronous Demodulator Signal Conditioning

The synchronous demodulator consists of a series and shunt analog transmission gate. The switching waveforms on the control gates are synchronized with the chopping control signals at both the S-Band modulator and predetection chopper. The input to the synchronous demodulators (both azimuth and elevation channels) is a 50 Hz modulation waveform with a peak voltage proportional to the predetection signal power. The synchronous demodulator essentially translates the 50 Hz modulation signal to dc with the amplitude information stored and held in capacitors C9 and C16. This information is then filtered in a low pass filter, $F_{3dB} = 100$ Hz (NBW = 160 Hz), to remove the high frequency noise components, and is then amplified in U10 and U11 to provide the system azimuth and elevation boresight error voltages (see Figure 6-8).

Signal gain for the module is 17 dB, and for noise 6 dB. For weak signal conditions, the output noise density will be ± 5 volts peak or ± 1.67 volts rms. Maximum boresight error voltages will be 316 millivolts and will drive to a minimum error voltage when a zero boresight error exists. The minimum voltage will be defined by the antenna precomparator null depth. For a null depth of 10 dB (output will drive to ± 100 millivolts), 20 dB (± 31.6 millivolts), 30 dB (± 10 millivolts), and for 40 dB (± 3.16 millivolts).

6.6.3 QUADRAPHASE TIMING

As mentioned previously, the quadrature timing and detection system are located in the same module. The timing scheme utilizes a dual monostable multivibrator operating in a stable mode for generation of 50 Hz pulse repetition rate with a 97.5 percent duty cycle. The pulse width for an On condition is 19.5 milliseconds and for the Off condition is 0.5 millisecond. This clock is then buffered and inverted to generate the correct phasing of the control signals to the 76T RF amplifier and to the quad analog switches. The effect of chopping is to generate a chopper-stabilized dc detector while being ac coupled to the final amplifier stage. Figure 6-7 shows the block diagram.

6.7 Environmental

Since all of the Monopulse Error Channel circuit devices and components are either similar or identical to those used in the NST circuit designs, no exception to the NST environmental specifications is necessary. The S-Band modulating switch and CMOS digital circuits are new, however, and will be required to meet both the temperature range and radiation environment.

The S-Band switch has an operating range of -60°C to $+110^\circ\text{C}$, but no radiation data was available at the time of report preparation. According to the "Radiation Design Criteria Handbook,"¹ which was initiated for the MJS '77 program, RCA CMOS circuits with a gate oxide annealing temperature of 950°C appeared to be quite adequate for

¹"Radiation Design Criteria Handbook," Jet Propulsion Laboratory Technical Memorandum 33-763, August 1, 1976.

radiation dose levels of less than $1.25 \times 10^{+5}$ rad (SI). For radiation dose levels in excess of $1.5 \times 10^{+5}$ rad (SI), CMOS devices will be required to have the radiation hard dry gate oxide process developed by RCA. The LM108AH operational amplifier will need to be hardened for all radiation dose levels, to minimize output offset voltage variations and input offset current variations.

Mission requirements will dictate what level of study will be required for the radiation environment.

SECTION 7

7. TEST RESULTS

7.1 TDM System

7.1.1 COMPARISON TEST RESULTS VERSUS SPECIFICATION

The table below presents a comparison of actual test results with the specification.

<u>Requirement</u>	<u>Specification</u>	<u>Test Results</u>
Sum channel threshold degradation	1/2 dB maximum	No discernible effect
Receiver best lock frequency change	$F_c \pm 1$ ppm	≈ 3 ppm
Command and ranging channel performance	No degradation	No effect
Angle channel SNR degradation	1 dB maximum	1.2 dB maximum
Angle channel crosstalk	-30 dB maximum	-55 dB maximum
Angle channel output variation	$\pm 5\%$ maximum	$\pm 5\%$, -50 to -140 dBm and -10 to +55°C. +5, -20% from -50 to -150 dBm and -10 to +55°C.
Angle channel output with no input (balance)	2% maximum of output level when sum input level = angle input level	1.7%
Amplitude slew rate	TBD	See Figure 7-14.

7.1.2 TDM PERFORMANCE DATA

7.1.2.1 Receiver Best-Lock Frequency

Figure 7-1 displays receiver best-lock frequency at 25°C for the angle channels On and Off. The load current of the angle channels drops the 9 and 6 volt power supplies by approximately 100 millivolts with a resulting 3-ppm change in best-lock frequency. The best-lock frequency is affected by a receiver $13F_1$ coherent spur which causes random variations and somewhat clouds the angle channel impact.

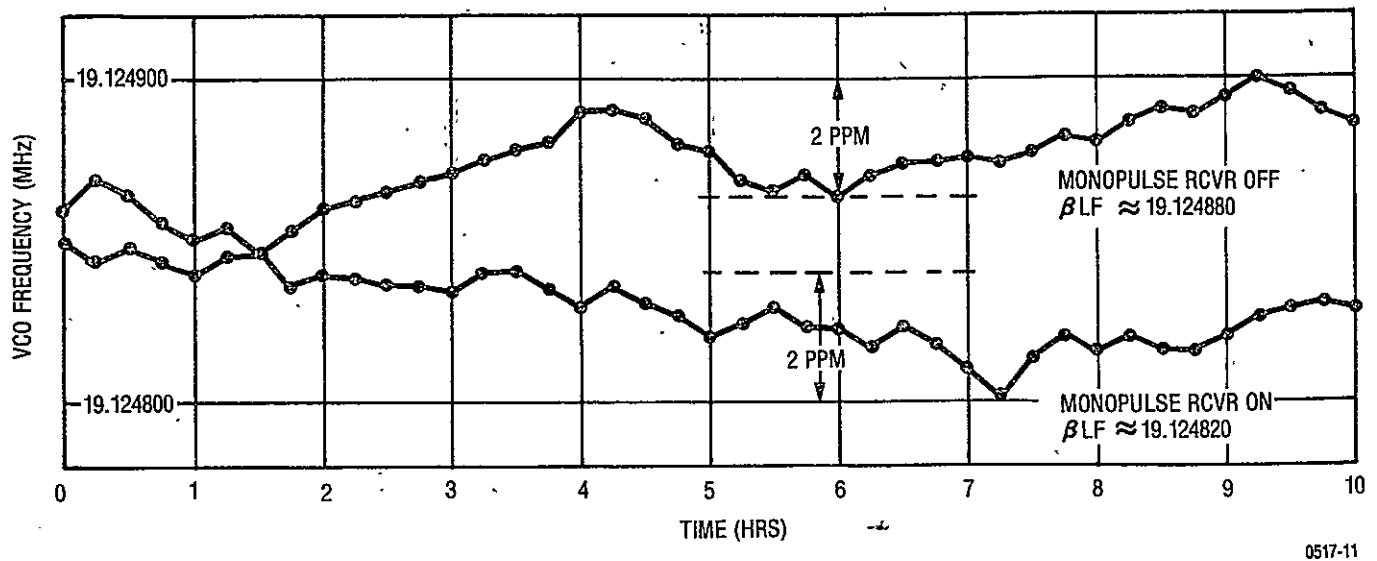


Figure 7-1. Uplink Frequency Uncertainty (20 Second Frequency Samples Averaged for 15 Min. Period at +25°C)

7.1.2.2 Command and Ranging Channel Performance

Figures 7-2 and 7-3 display command and ranging channel signal and noise levels as a function of input level. There is no discernible effect when the angle channels are energized.

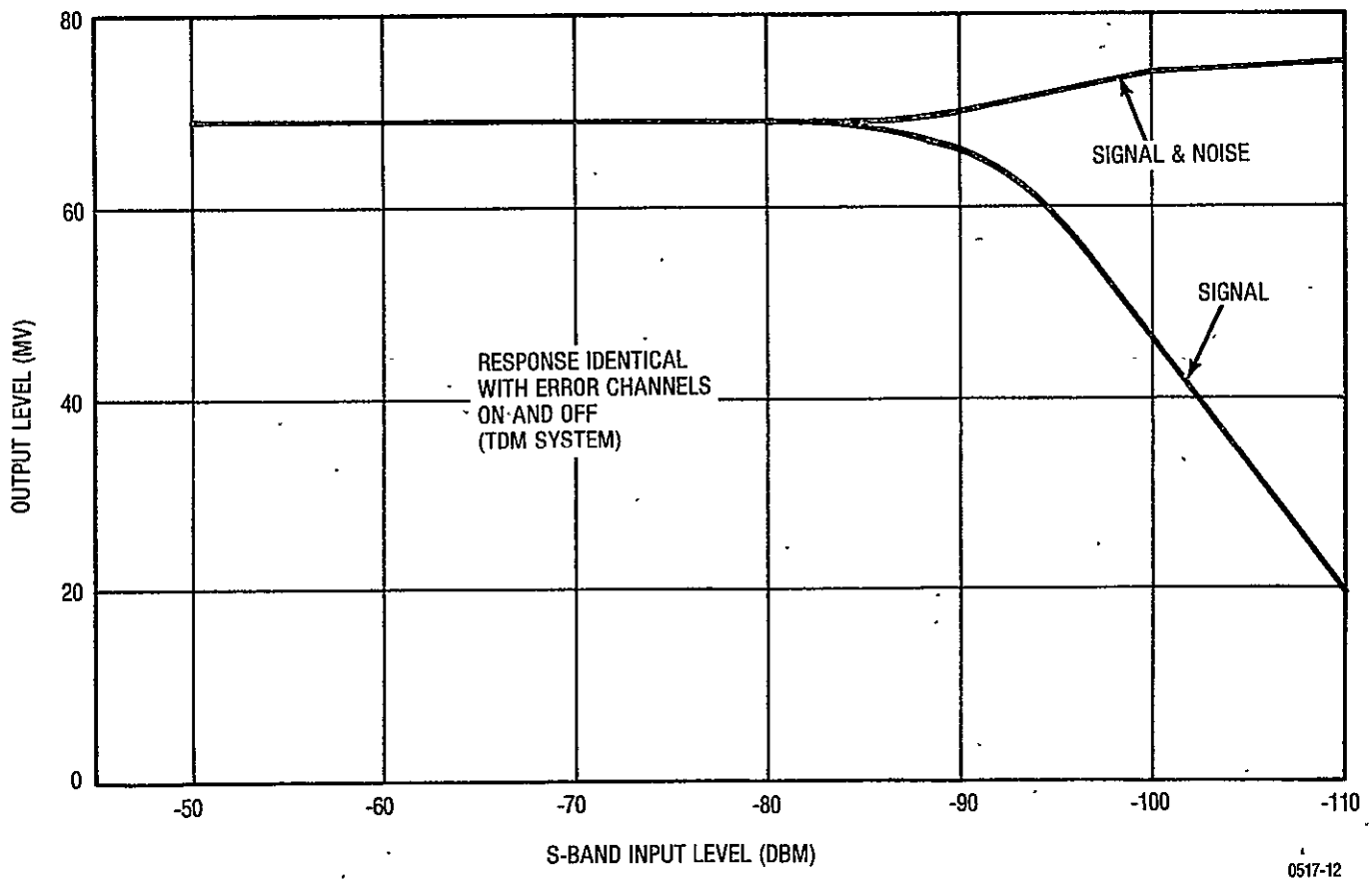


Figure 7-2. Micromin Ranging Channel Output vs RF Input (Temp: +25°C; Mod Index: 1 Rad; Mod Tone: 500 kHz)

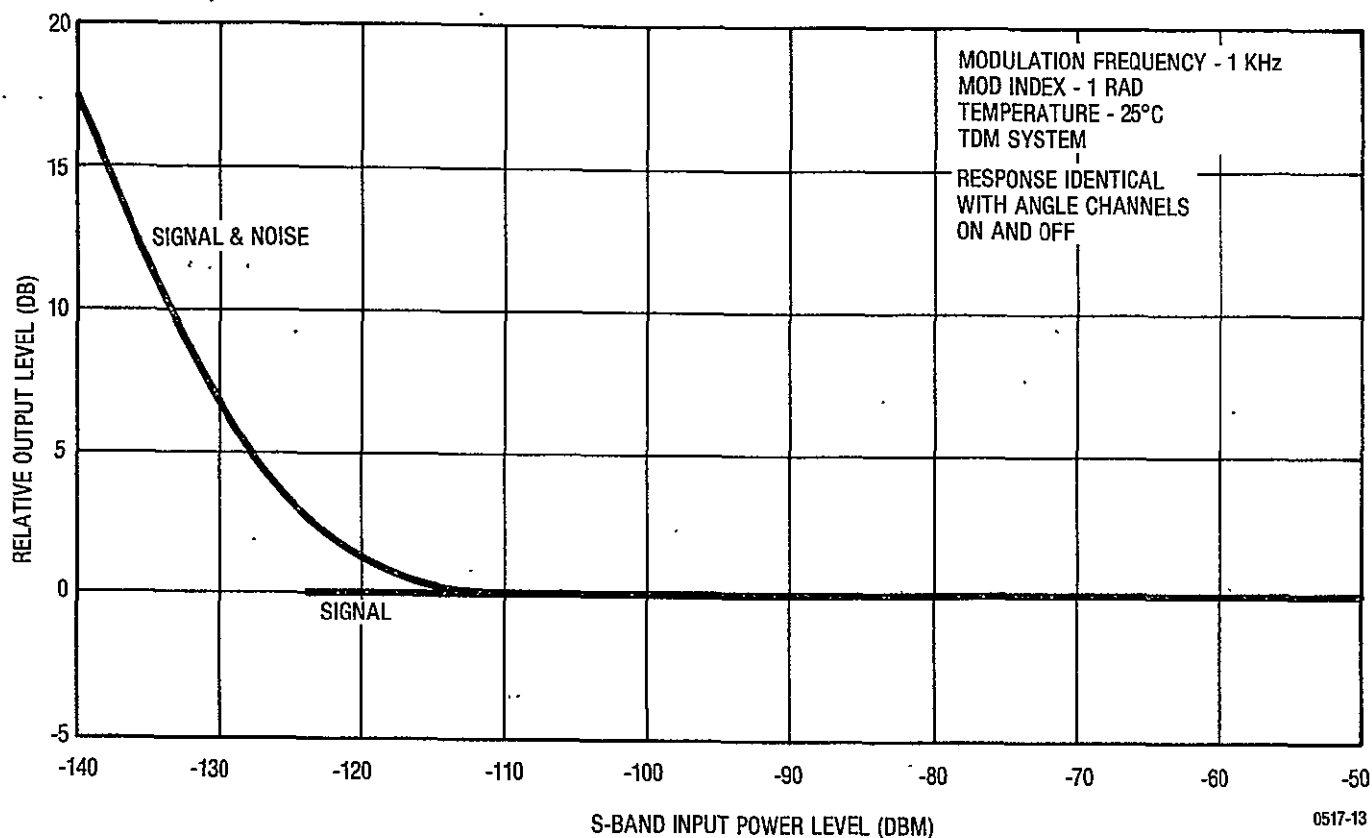


Figure 7-3. Micromin Command Output vs S-Band Signal Level (Modulation Frequency: 1 kHz; Mod Index: 1 Rad; Temperature: ~25°C)

7.1.2.3 Angle Channel Signal-to-Noise Ratio Degradation from Theoretical

Table 7-1 displays the angle channel signal-to-noise ratio for various Σ/Δ level ratios and the angle channel input level at -138 dBm which is the theoretical unity signal to noise ratio level. The variations with temperature are primarily due to receiver noise figure variation. The average value of signal-to-noise ratios is about 0.7 dB which is primarily determined by the insertion loss of S-Band switch.

Table 7-1. TDM Signal-to-Noise Ratio (At Angle Channel Level = -138 dBm)

Channel	Σ/Δ (dB)	Temperature		
		-10°C	+25°C	+55°C
X	10	-0.6	-0.8	-0.8
	20	0.0	-1.0	-0.9
	30	-0.5	-1.1	-1.0
Y	10	-0.5	-0.6	-0.8
	20	0.0	-0.7	-0.8
	30	-0.2	-0.7	-1.2

Table 7-1. Micromin Command Output vs S-Band Signal Level (Modulation Frequency: 1 kHz; Mod Index: 1 Rad; Temperature: ~25°C) (Contd)

KTB	=	-174 dBm/Hz
NF	=	8 DB
NBW	=	25 dB (2Bn = 320 Hz)
Two-channel loss	=	3 dB
Theoretical units S/N	=	-138 dBm

7.1.2.4 Angle Channel Crosstalk

Table 7-2 presents the angle channel crosstalk performance.

Table 7-2. Crosstalk Performance

Crosstalk	Temperature		
	-10°C	+25°C	+55°C
X - Y	-61.1 dB	-64.6 dB	-55 dB
Y - X	-56.6	-64.7	-60

7.1.2.5 Angle Channel Output Variation

Figures 7-4 and 7-5 display the azimuth and elevation channel output level variations with temperature and signal level. The system meets the requirements over the -10 to +55°C temperature range from -50 to -140 dBm. From -140 to -150 dBm the output level varies to -20 percent of expected value. The variations near threshold are attributed to a coherent spur that is known to exist in the microminiature receiver. The interference level is equivalent to a -175 dBm signal which is the approximate estimated level of the spur.

Figures 7-6 and 7-7 display the same information for the angle channels phased to provide a negative output.

Figures 7-8 and 7-9 display angle channel outputs with ± 150 kHz Doppler applied.

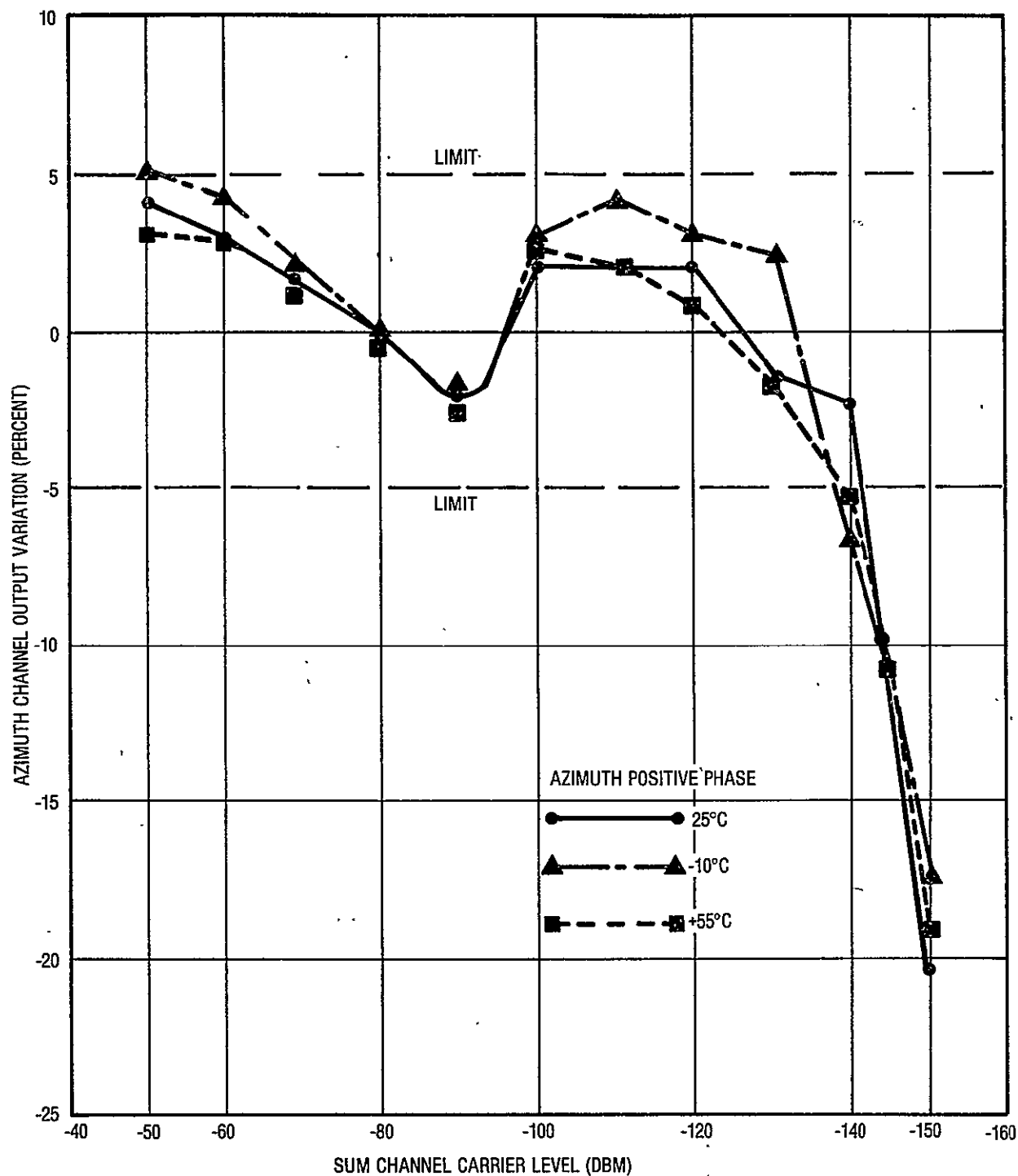
Figures 7-10 and 7-11 show the angle channel output level as a function of sum-to-angle channel ratios (Σ/Δ), which demonstrates system linearity.

7.1.2.6 Angle Channel Output with No Input

Figures 7-12 and 7-13 display the dc output of the angle channels as a function of signal level and temperature. It is seen that the offsets remain within the 2 percent limit. Level measurements below -135 dBm were found to vary slowly with time (i.e. 10 minutes for peak-to-peak excursions) which again indicates the presence of a coherent spurious signal.

7.1.2.7 Amplitude Slew Rate

Figure 7-14 displays the angle channel output variation as the S-Band input is swept 10 dB at a 30 dB per second rate. A positive-going modulating voltage corresponds to a decreasing S-Band level and a consequent decreasing angle channel voltage. The output voltage waveform is cusped because the input is modulated linearly in dB per volt.



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Figure 7-4. TDM Angle Channel Level Variation vs Signal Level and Temperature (Azimuth: Positive Phase)

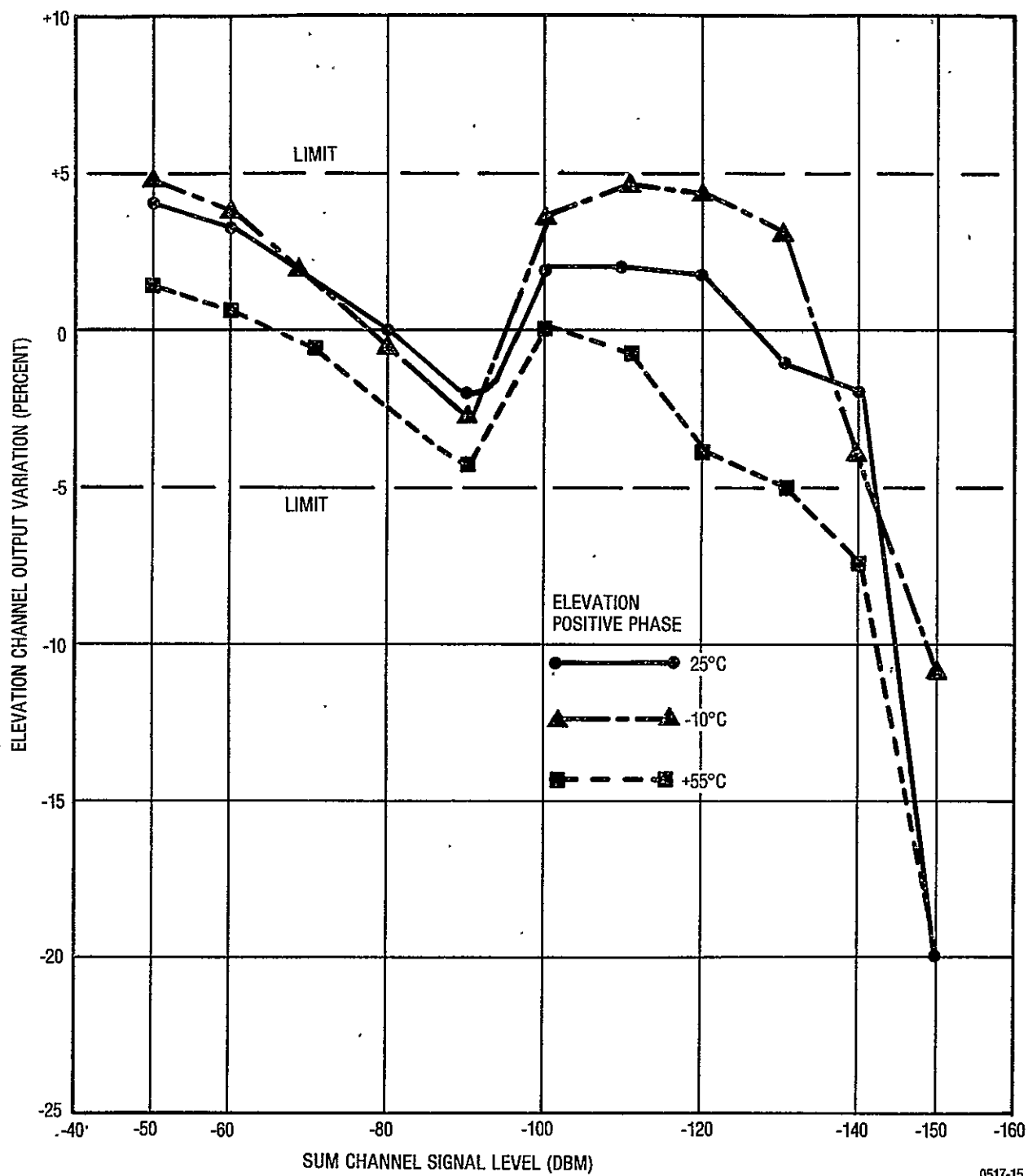


Figure 7-5. TDM Angle Channel Level Variation vs Signal Level and Temperature (Elevation: Positive Phase)

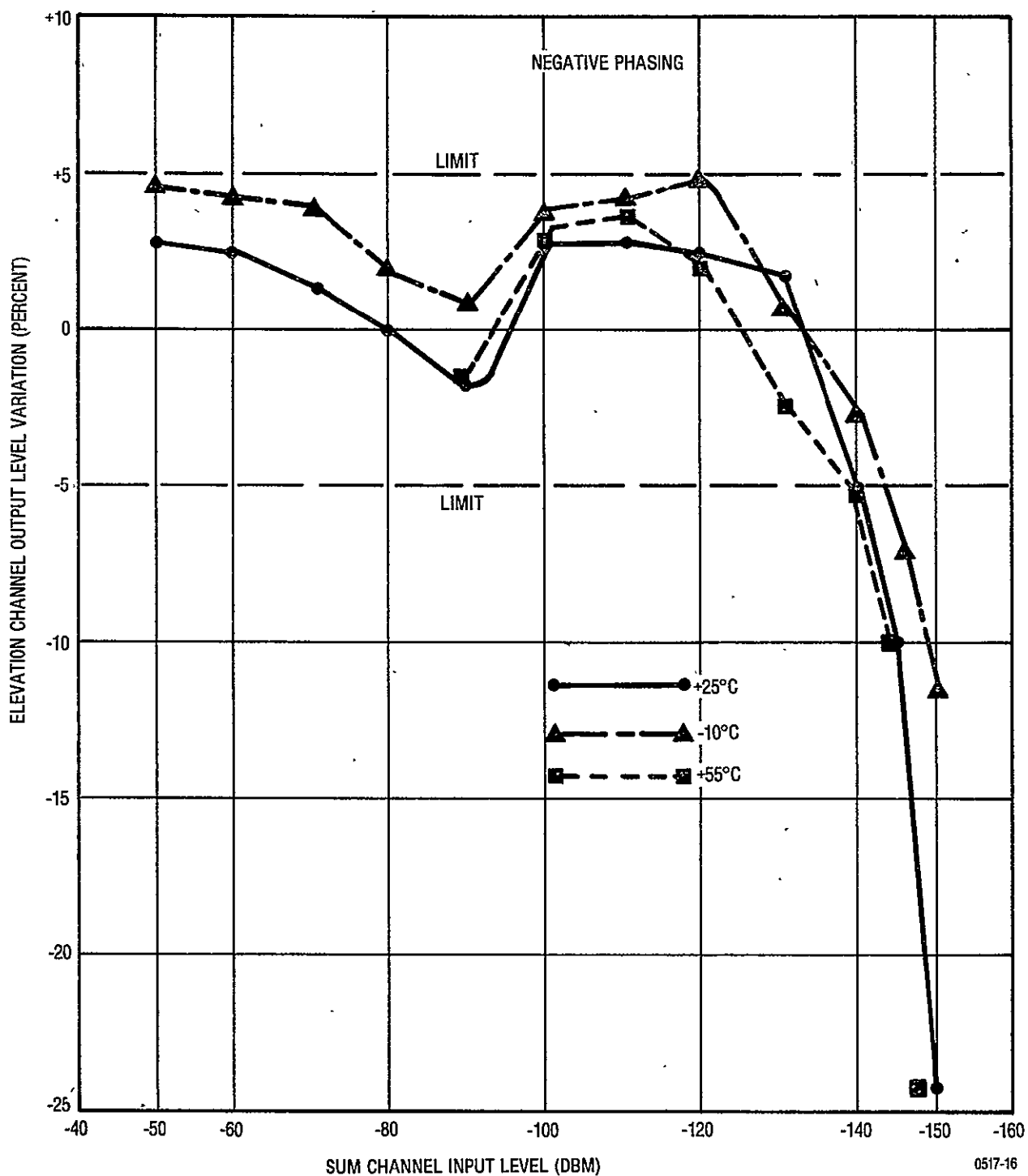


Figure 7-6. TDM Elevation Channel Output Variation vs Signal Level and Temperature (Negative Phasing)

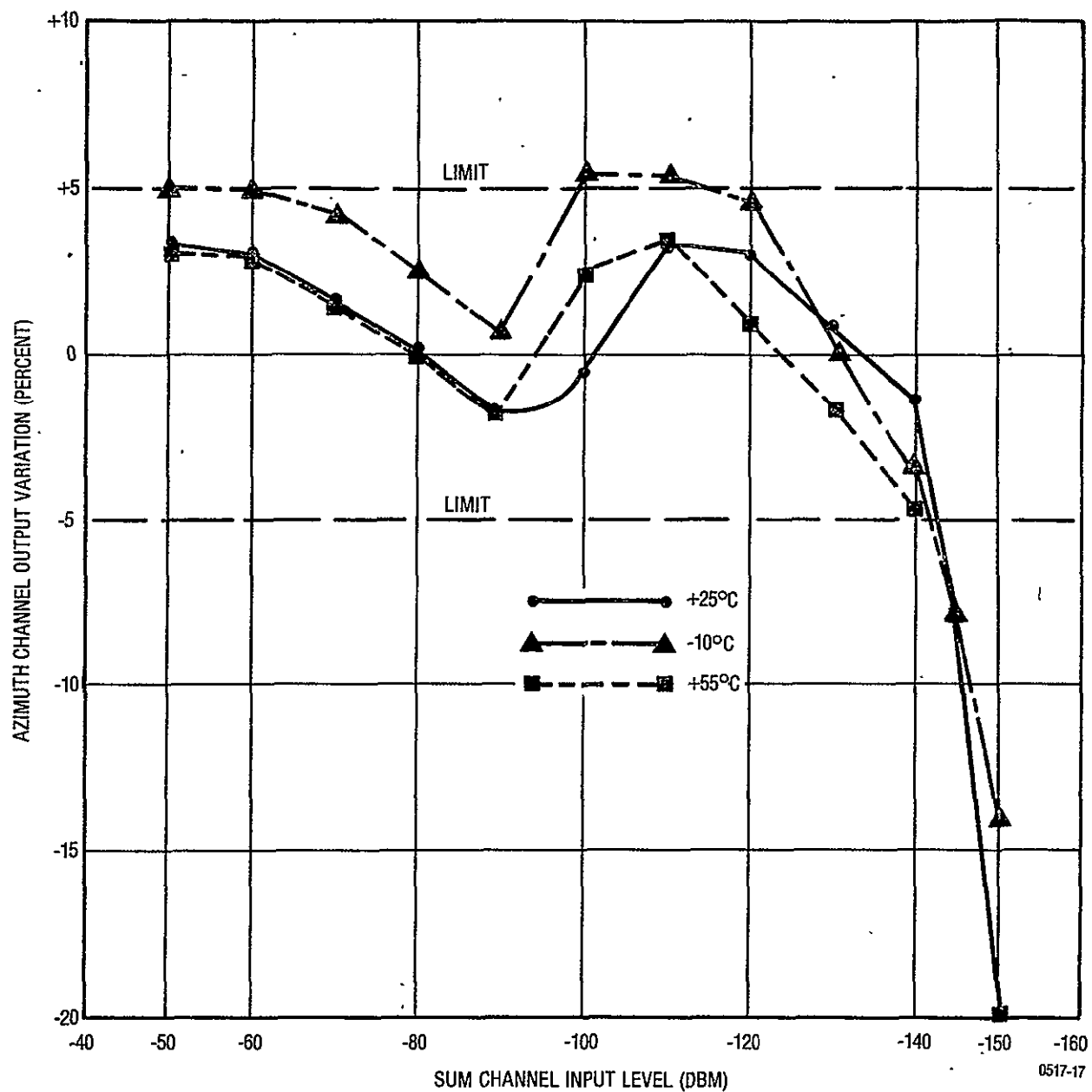


Figure 7-7. TDM Azimuth Channel Level Variation vs Signal Level and Temperature (Negative Phasing)

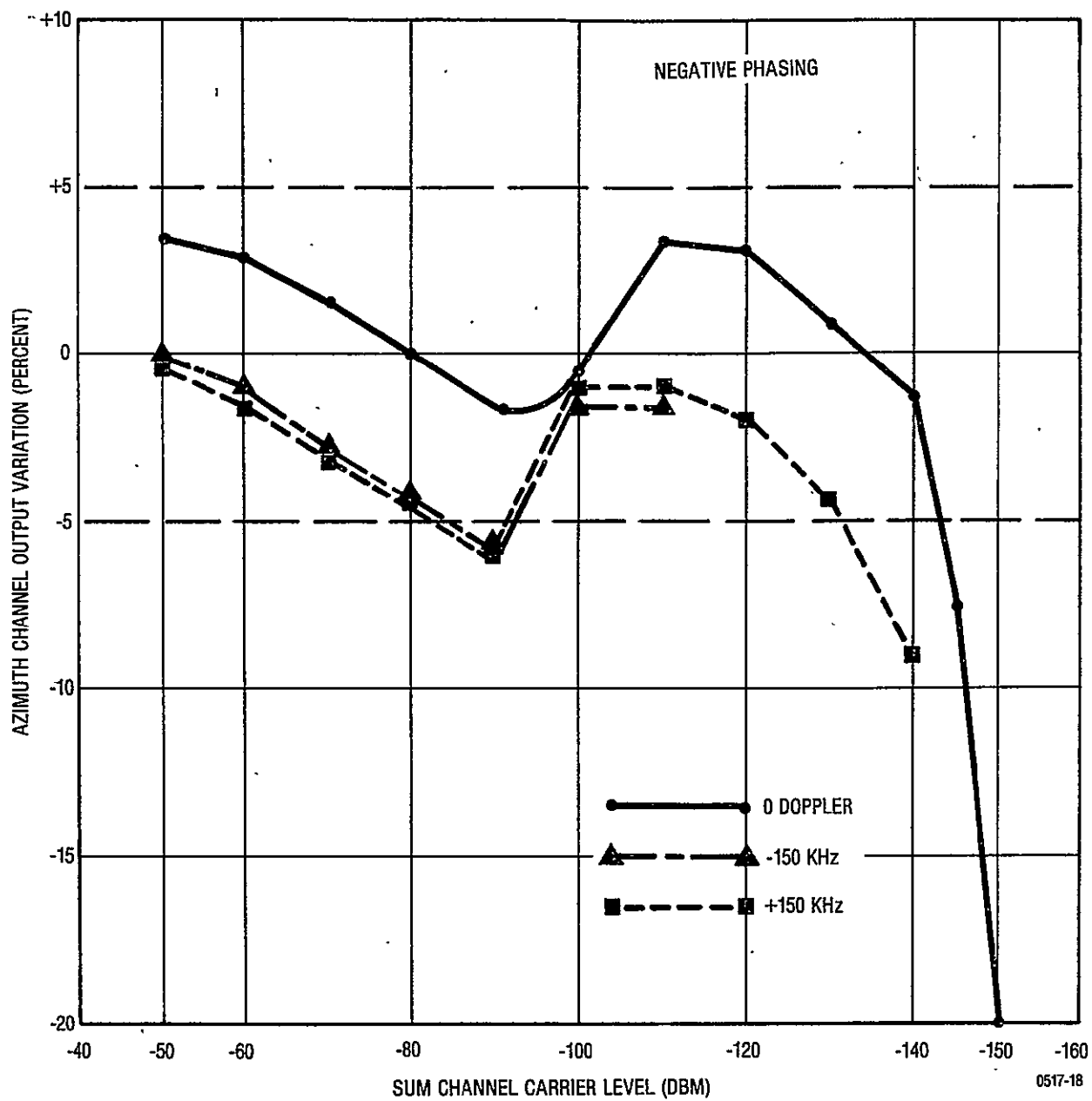


Figure 7-8. TDM Azimuth Channel Output Variation with S-Band Level and Doppler (Negative Phasing)

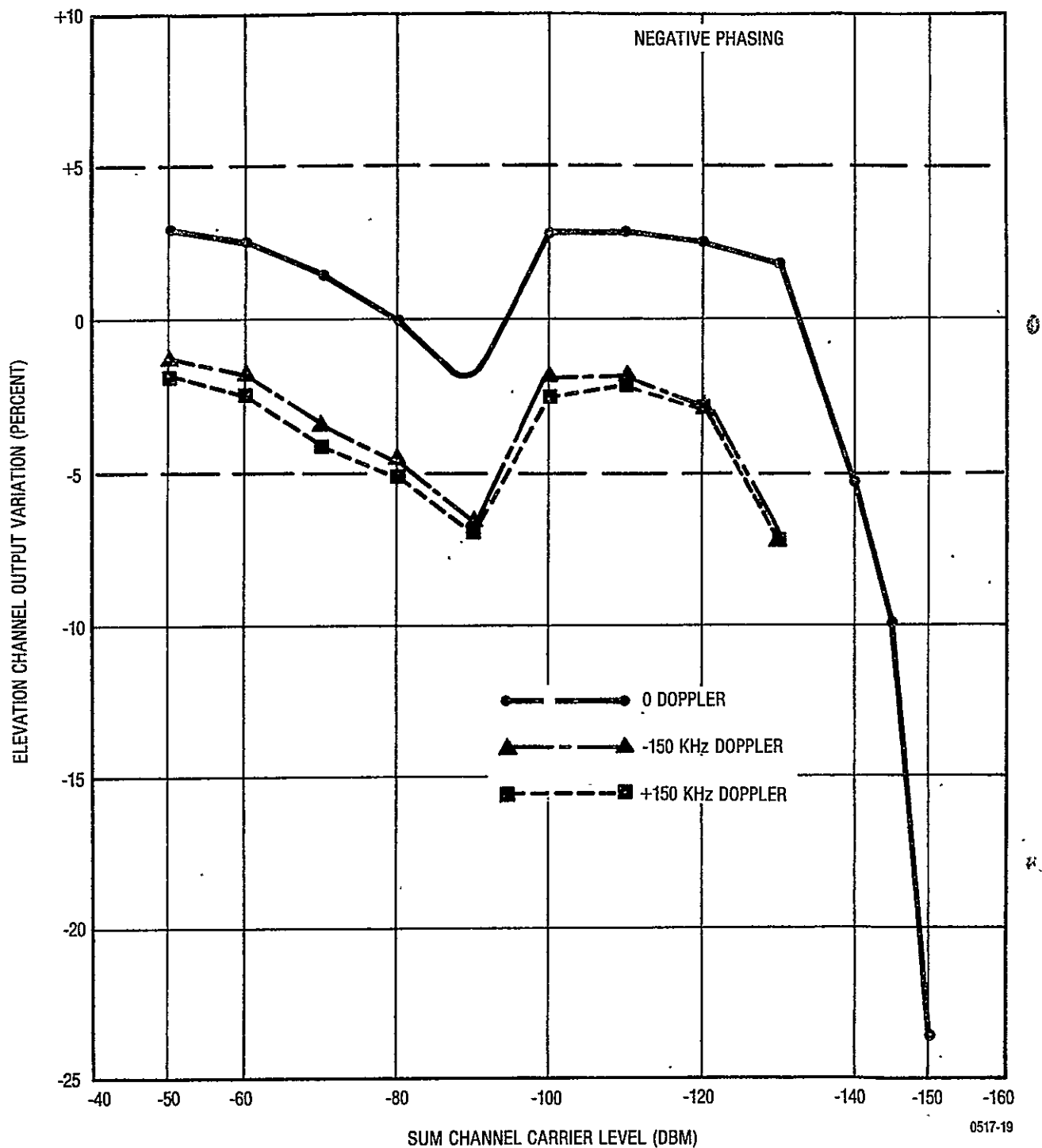


Figure 7-9. TDM Elevation Channel Output Variation with S-Band Level and Doppler (Negative Phasing)

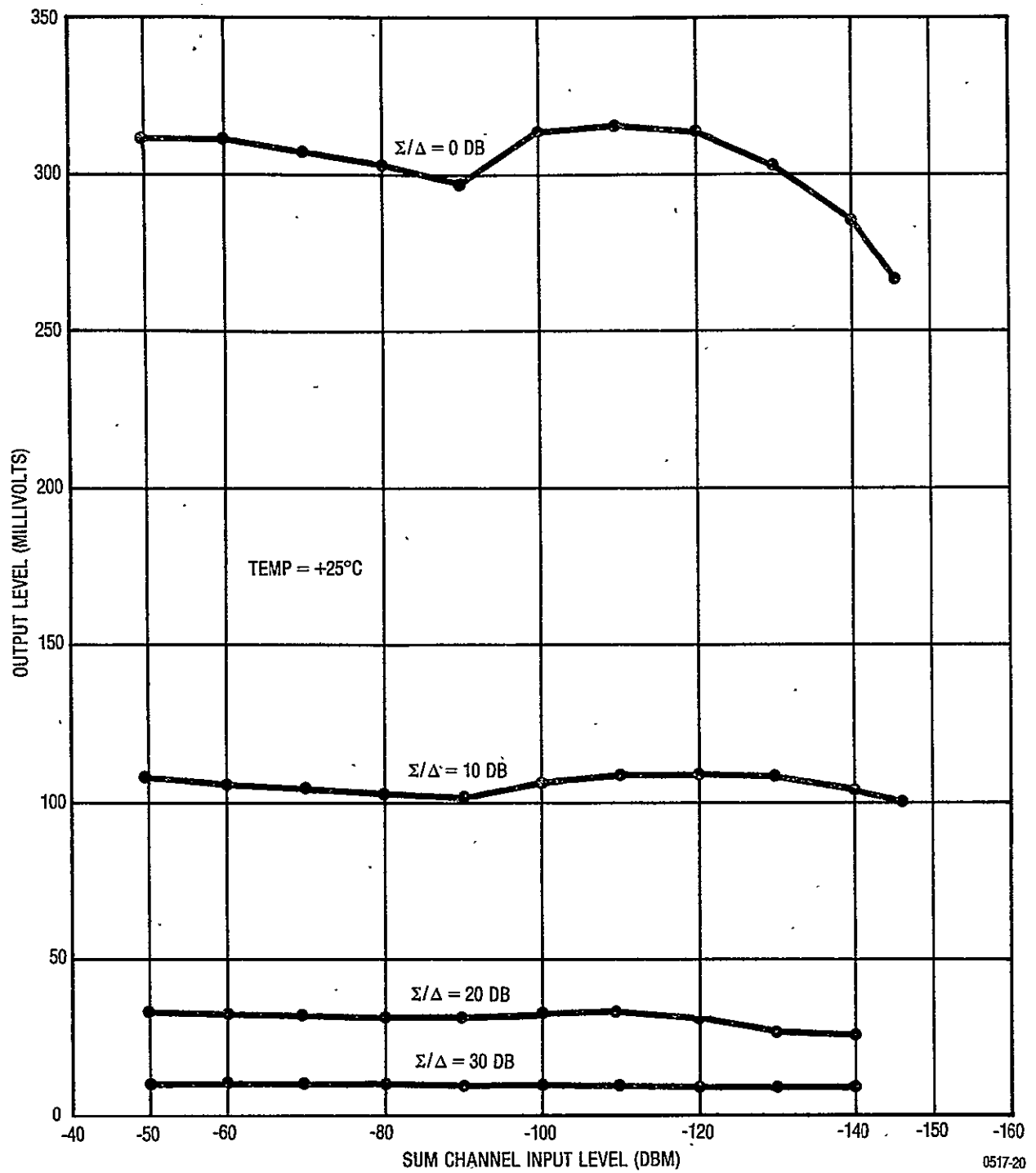


Figure 7-10. Monopulse Receiver (TDM & QPH) Angle Channel Output Level vs Input Level (Temperature = + 25°C)

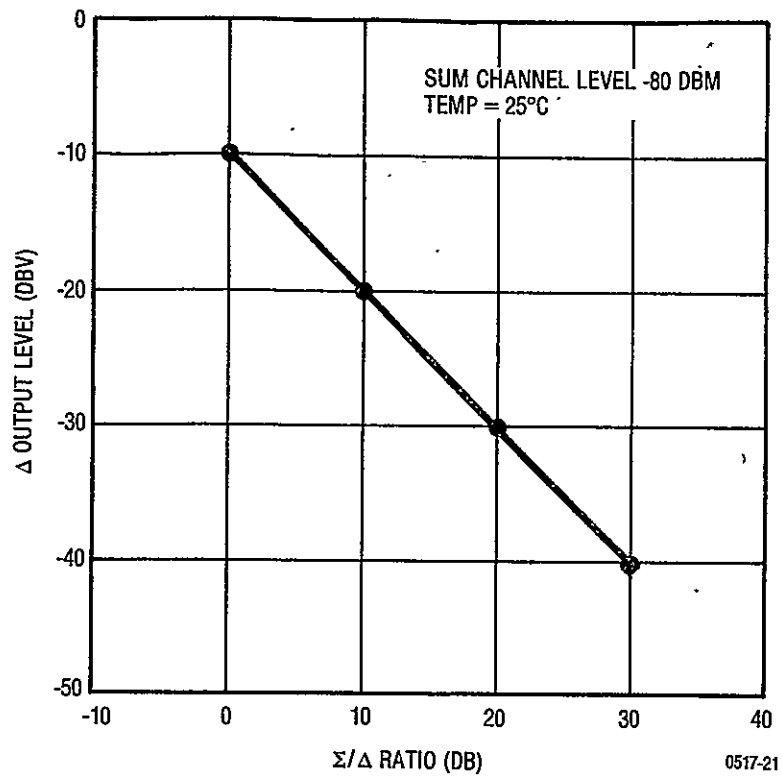


Figure 7-11. Monopulse (TDM & QPH) Angle Channel Output Level vs Sum-To-Angle Signal Level Ratio

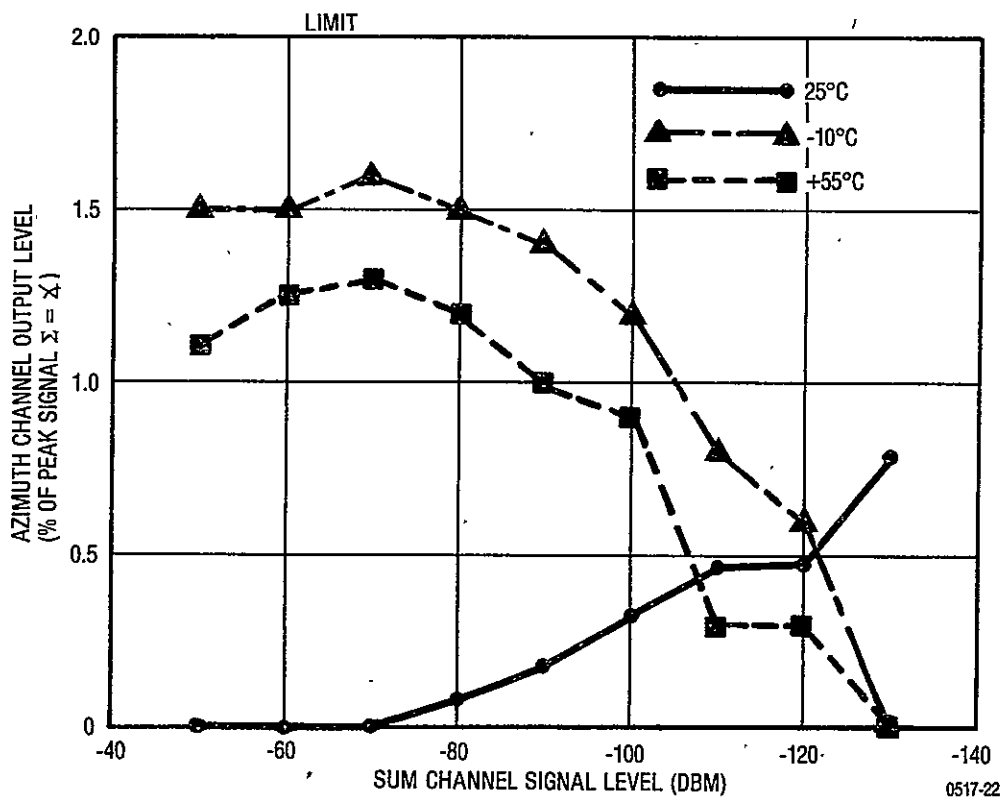


Figure 7-12. TDM Azimuth Channel Balance vs Signal Level and Temperature

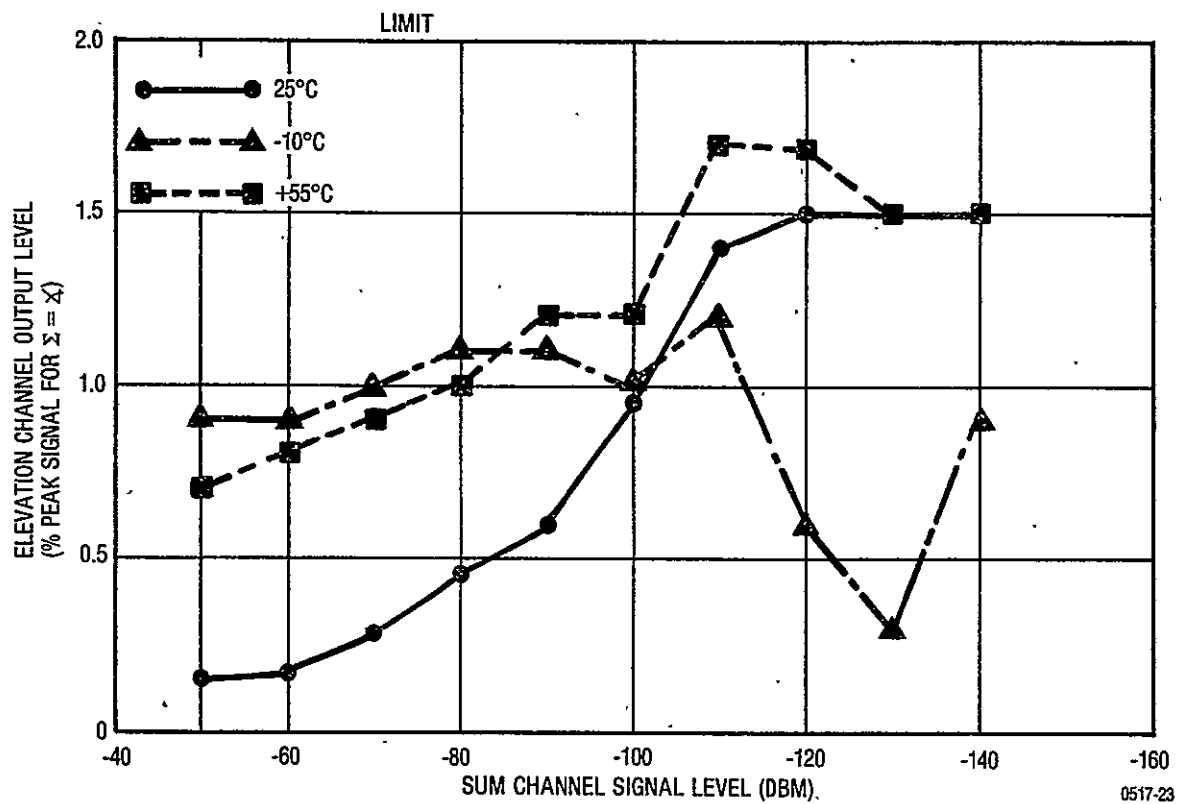


Figure 7-13. TDM Elevation Channel Balance vs Signal Level and Temperature

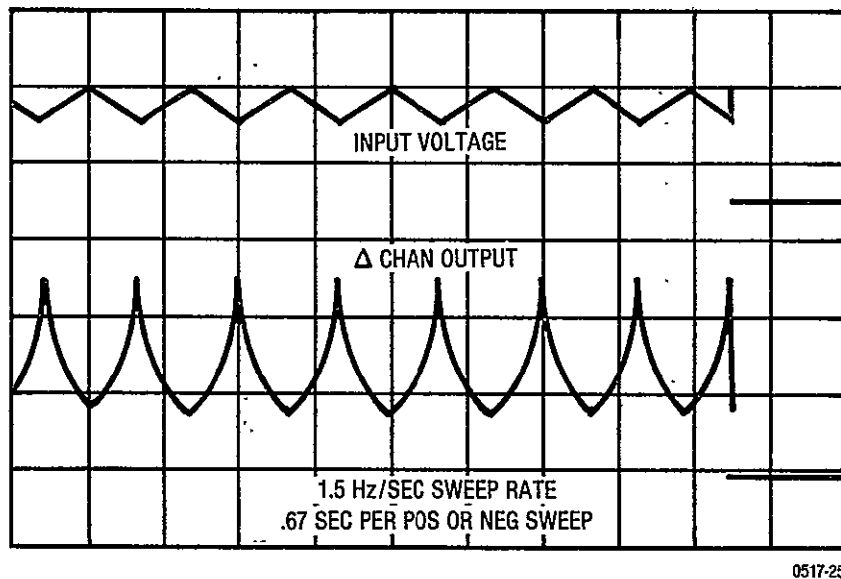


Figure 7-14. TDM Monopulse Receiver Response to 30 dB/second Angle Channel Slew Rate (Total Input Sweep 10 dB at 30 dB/second)

It is seen that the output amplitude response lags the input signal approximately 1 dB. However, the turn around response to the input amplitude reversals is within the 10-millisecond response time of the sampling system.

7.2 Quadrature System

7.2.1 TEST RESULTS VERSUS SPECIFICATION

The table below presents a comparison of actual test results with the specification.

<u>Requirement</u>	<u>Specification</u>	<u>Test Results</u>
Sum channel threshold degradation	1/2 dB maximum	No discernible effect
Receiver best lock frequency	$F_c \pm 1$ ppm	≈ 3 ppm
Command and ranging channel performance	No degradation	No effect
Angle channel SNR degradation	1 dB maximum	1.6 dB
Angle channel crosstalk	-30 dBm maximum	-15 dB worst case
Angle channel output variation	$\pm 5\%$	$\approx \pm 20\%$
Angle channel output with no input (balance)	2% maximum of output level when sum channel input level = angle channel input level	2% maximum
Amplitude slew rate	TBD	See Figure 7-14

7.2.2 QUADRATURE PERFORMANCE DATA

7.2.2.1 Receiver Best Lock Frequency

The effects on best lock frequency are the same as the effects of the TDM System.

7.2.2.2 Command and Ranging Channel Performance

The Quadrature System has no effect on the sum channel command and ranging performance.

7.2.2.3 Angle Channel Signal to Noise Ratio Degradation

Table 7-3 displays the angle channel signal to noise ratios for various Σ/Δ ratios and an angle channel input of -138 dBm which is the theoretical unity signal to noise ratio level. The average signal to noise ratio degradation is 1.1 dB for the Azimuth and 0.7 dB for the elevation. This is due to amplitude unbalances of the input coupling system.

Table 7-3. Quadrphase Signal-to-Noise Ratio (At Angle Level = -138 dBm)

Channel	Σ/Δ (dBm)	Temperature		
		-10°C	+25°C	+55°C
X	10	-0.9	-1.05	-1.6
	20	-0.9	-1.2	-1.4
	30	-1.2	-1.3	-1.6
Y	10	-0.3	-0.5	-1.06
	20	-0.7	-0.3	-1.2
	30	-0.7	-0.6	-0.72
KTB = -174 dBm/Hz NF = 8 dB NBW = 3 dB Two channel loss = 3 dB Theoretical unity S/N = -138 dBm				

7.2.2.4 Angle Channel Crosstalk

Table 7-4 displays the angle channel crosstalk performance. The worst case performance of 15 to 16 dB is as predicted.

Table 7-4. Crosstalk Performance

Crosstalk	Temperature		
	-10°C	+25°C	+55°C
X - Y	-16 dB	-55 dB	-18.5 dB
Y - X	-22 dB	-47 dB	-15.6 dB

7.2.2.5 Angle Channel Output Variation

Figure 7-15 displays the azimuth channel output as a function of S-Band input level and temperature. The change in output level is seen to be ± 20 percent from -50 to -140 dBm and over the temperature range. This is about four times worse than the TDM System performance. Figure 7-16 shows the azimuth output with the elevation channel input terminated. The performance then compares with TDM performance and indicates that crosstalk is affecting the performance. Figures 7-17, 7-18, and 7-19 display this more graphically, showing azimuth output (X) for both X and Y signals in, X-only in and Y-only in. The Y-only-in curve is the direct measure of crosstalk. Figure 7-20 shows this crosstalk component plotted as phase shift rather than voltage indicating approximately 10 degrees of phase shift as a function of temperature. Figure 7-21 indicates performance with ± 120 kHz of Doppler.

7.2.2.6 Angle Channel Output with No Input

Figure 7-22 displays the dc output offset voltage as a function of temperature and signal level. It is seen that the balance just meets the 2 percent requirement. Figures 7-23 through 7-29 display the same data for the elevation channel output. Results are essentially the same.

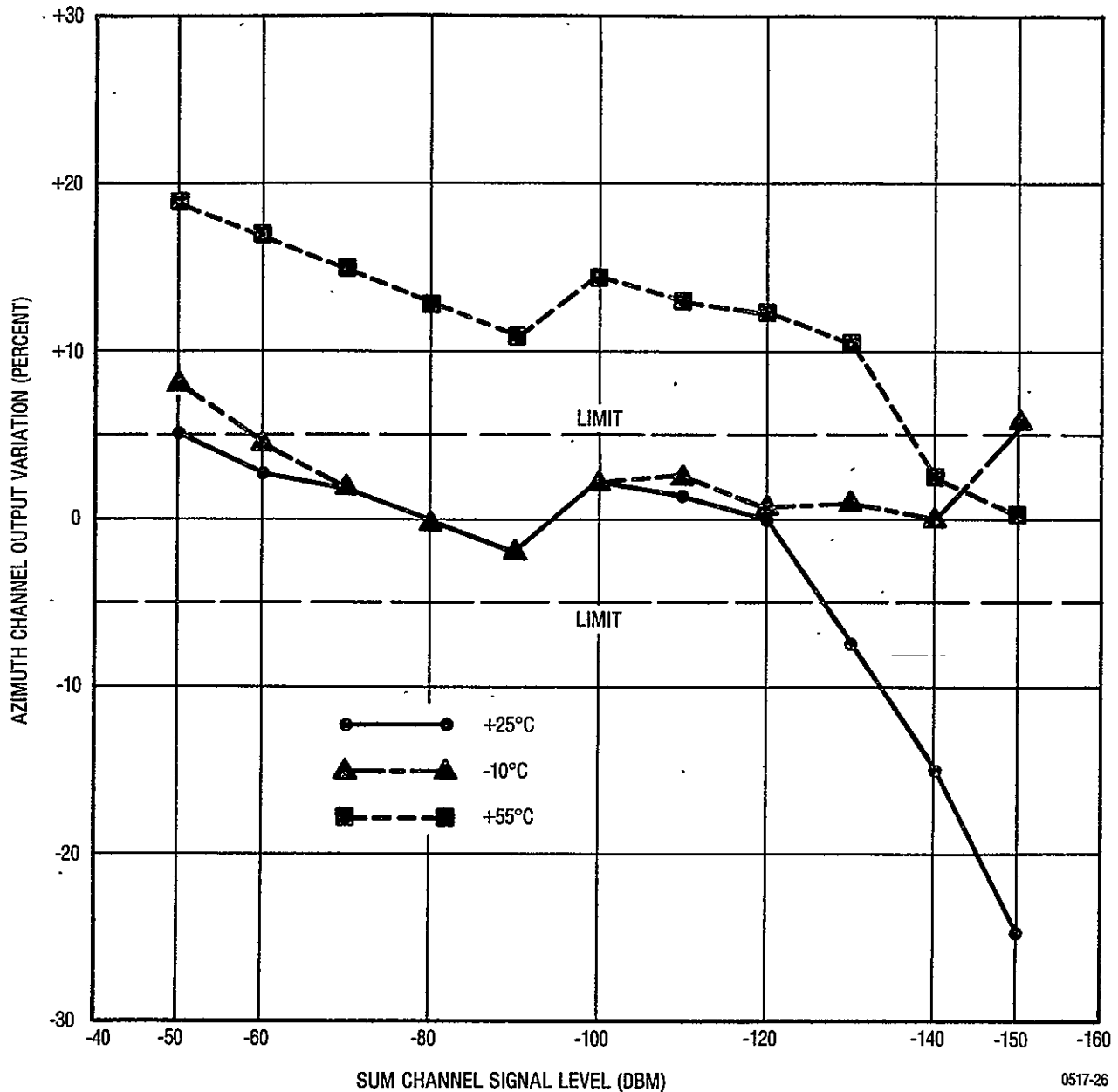


Figure 7-15. Quadrphase Azimuth Channel Output Amplitude vs Signal Level and Temperature

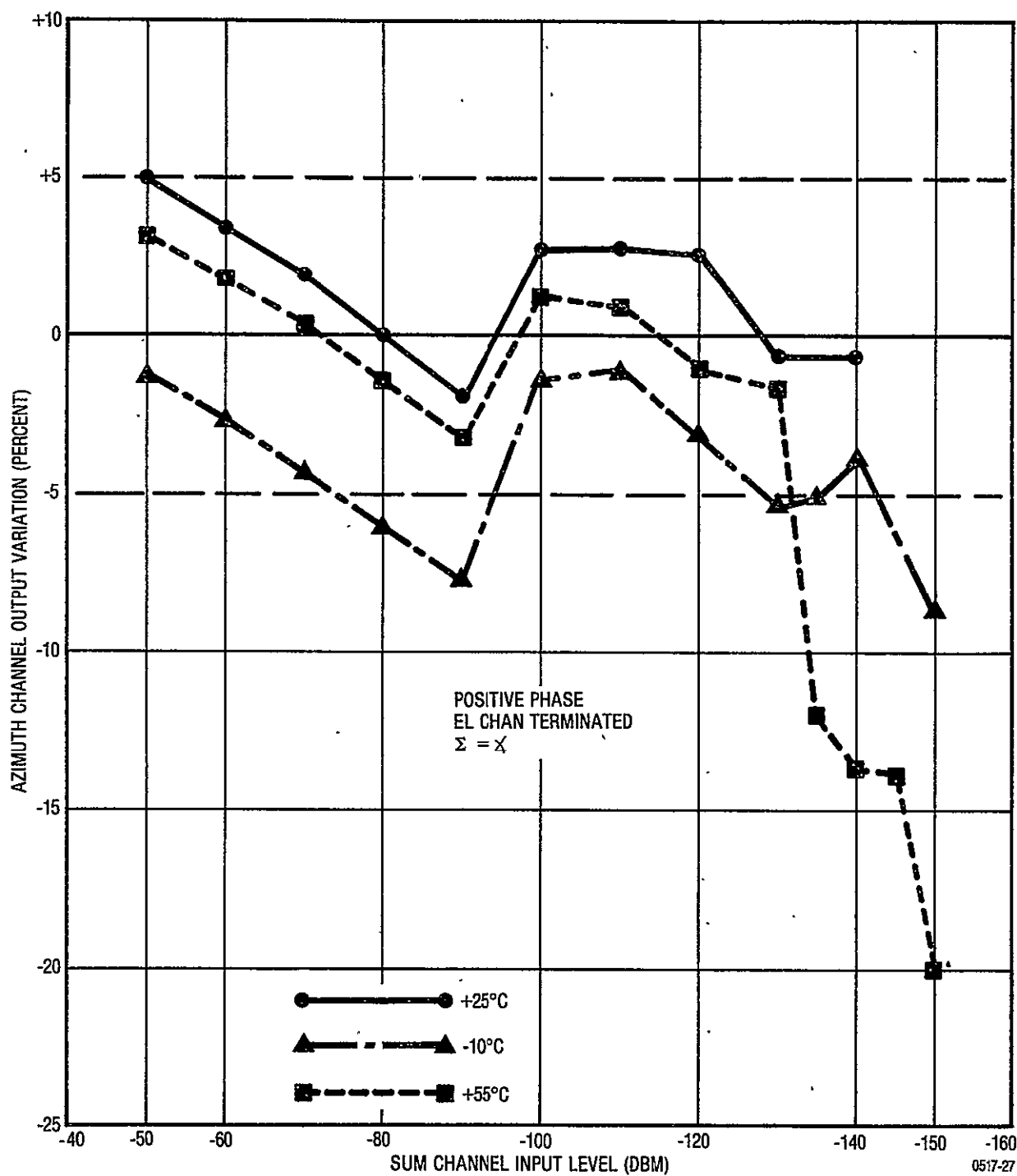


Figure 7-16. Quadraphase Azimuth Channel Level Variation vs Signal Level and Temperature

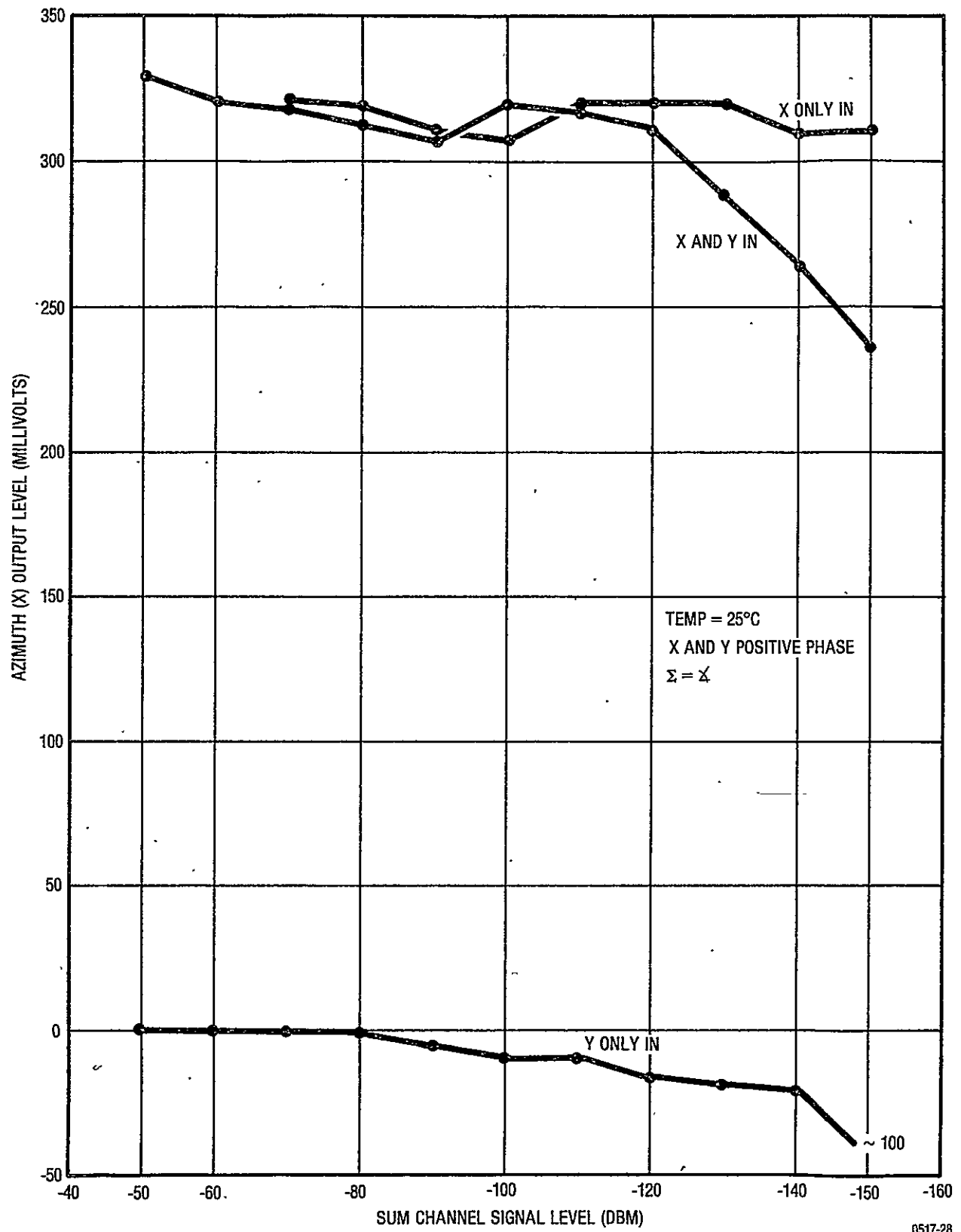


Figure 7-17. Quadrphase Azimuth Channel Output Level for Various Inputs

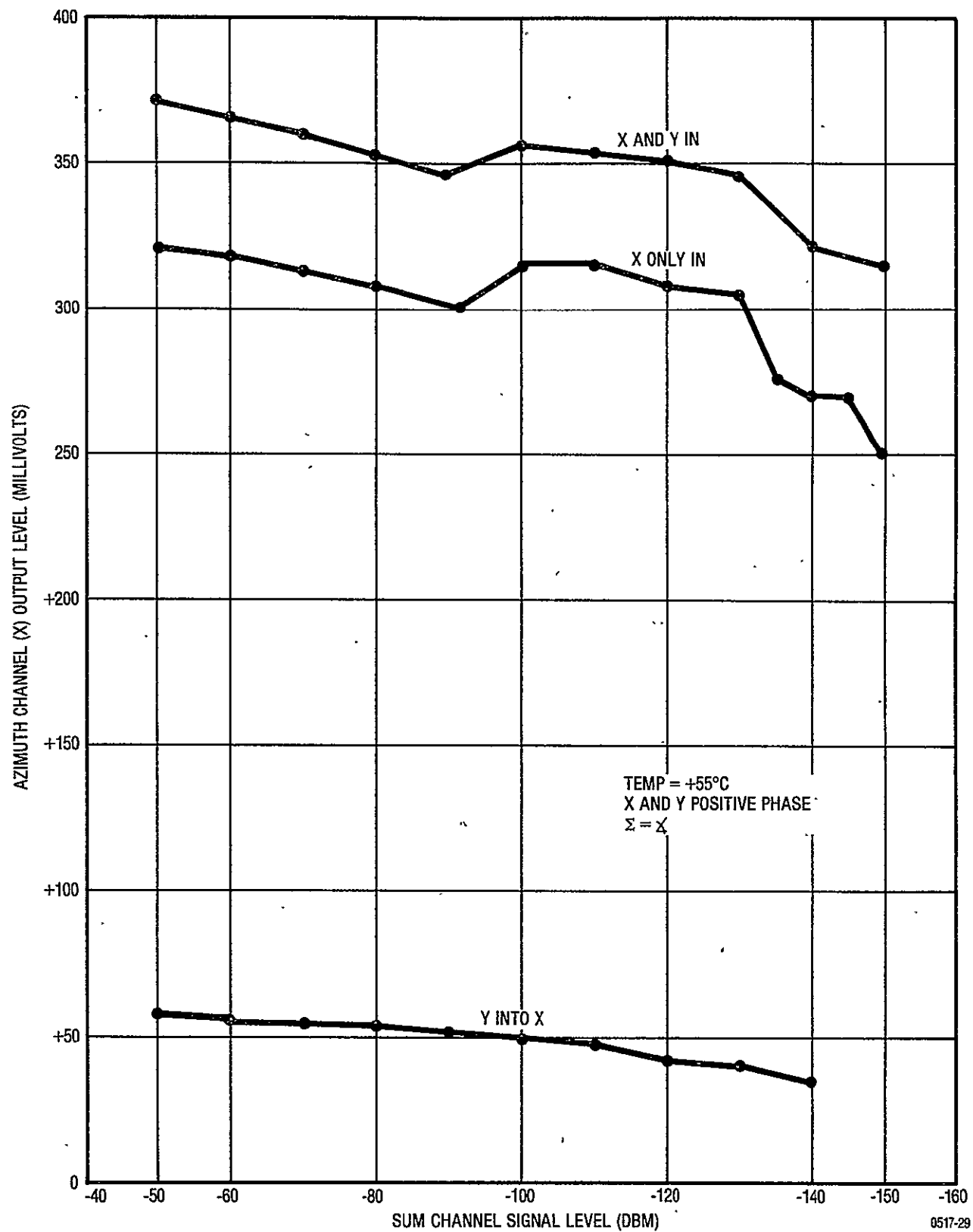


Figure 7-18. Quadraphase Azimuth Channel Output Level for Various Inputs

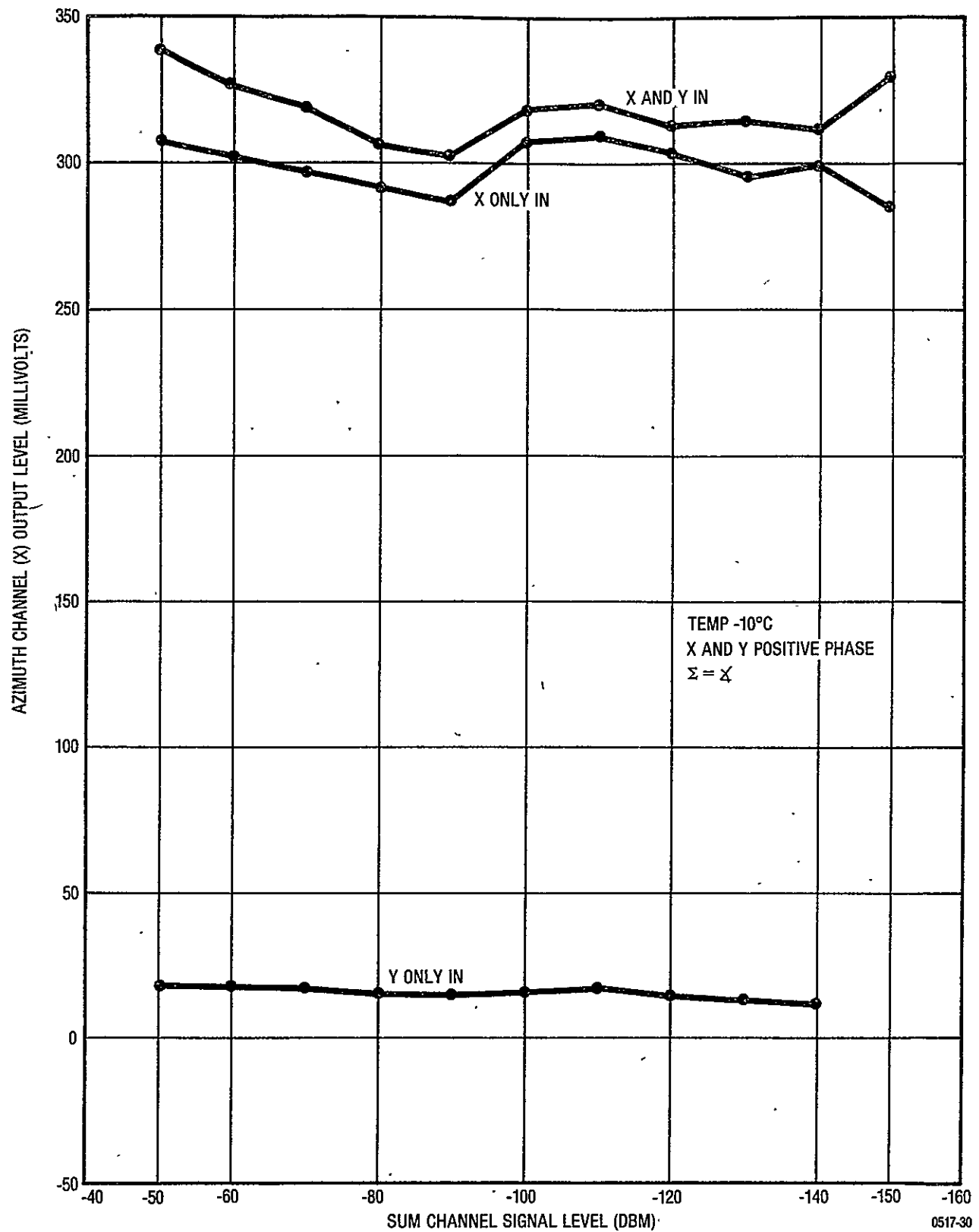


Figure 7-19. Quadrphase Azimuth Channel Output Level for Various Inputs

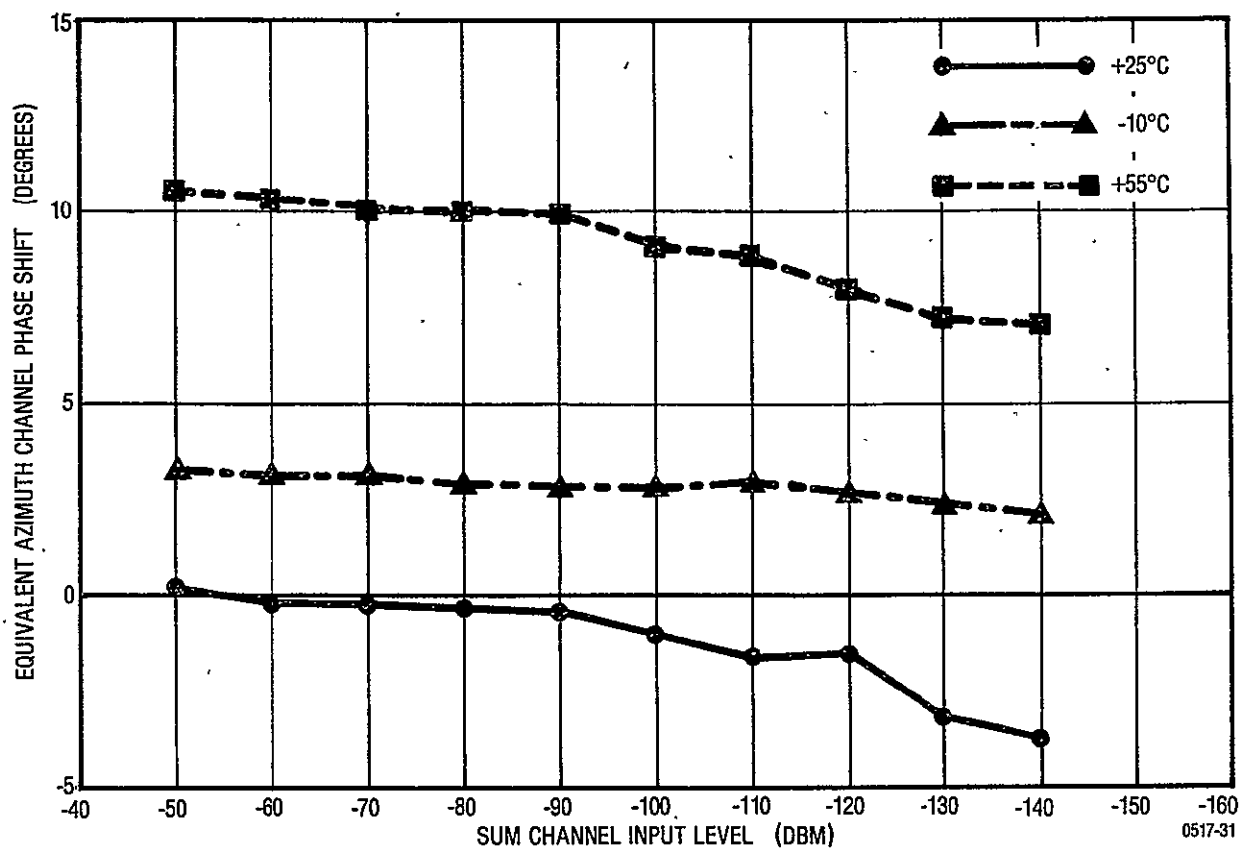


Figure 7-20. Quadrphase Azimuth Channel Phase Shift vs Input Level and Temperature

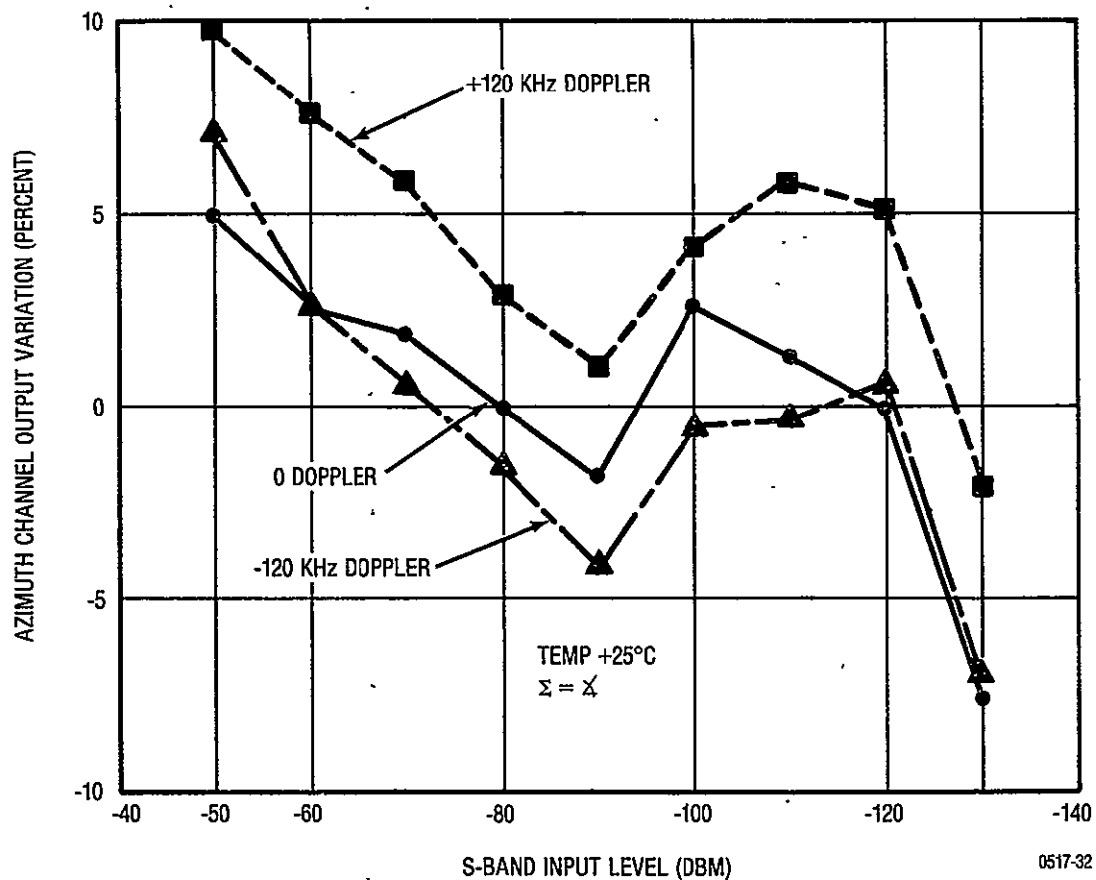


Figure 7-21. Quadrphase Azimuth Channel Output Variation with Doppler

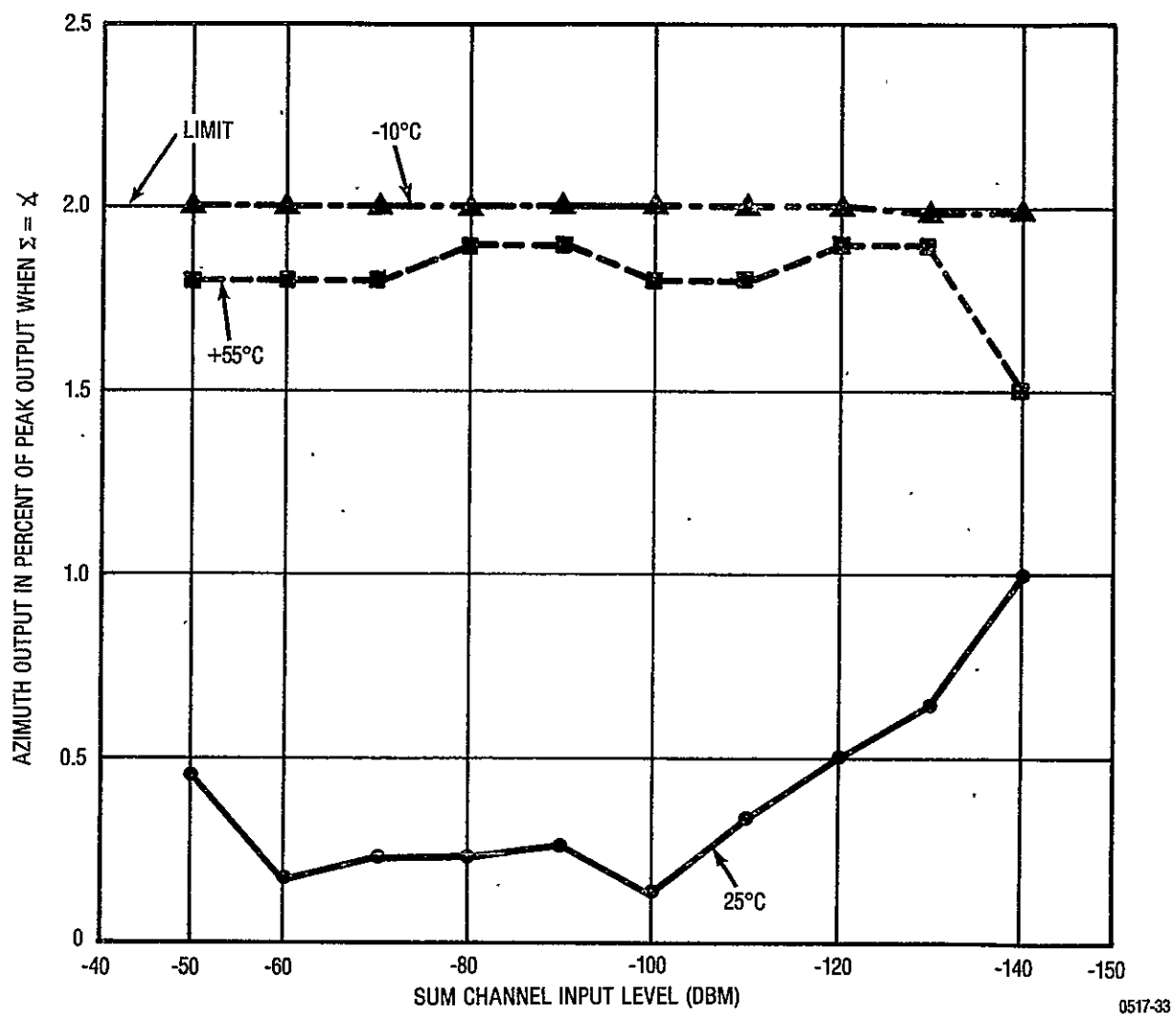


Figure 7-22. Quadrphase Azimuth Channel Balance vs Signal Level and Temperature

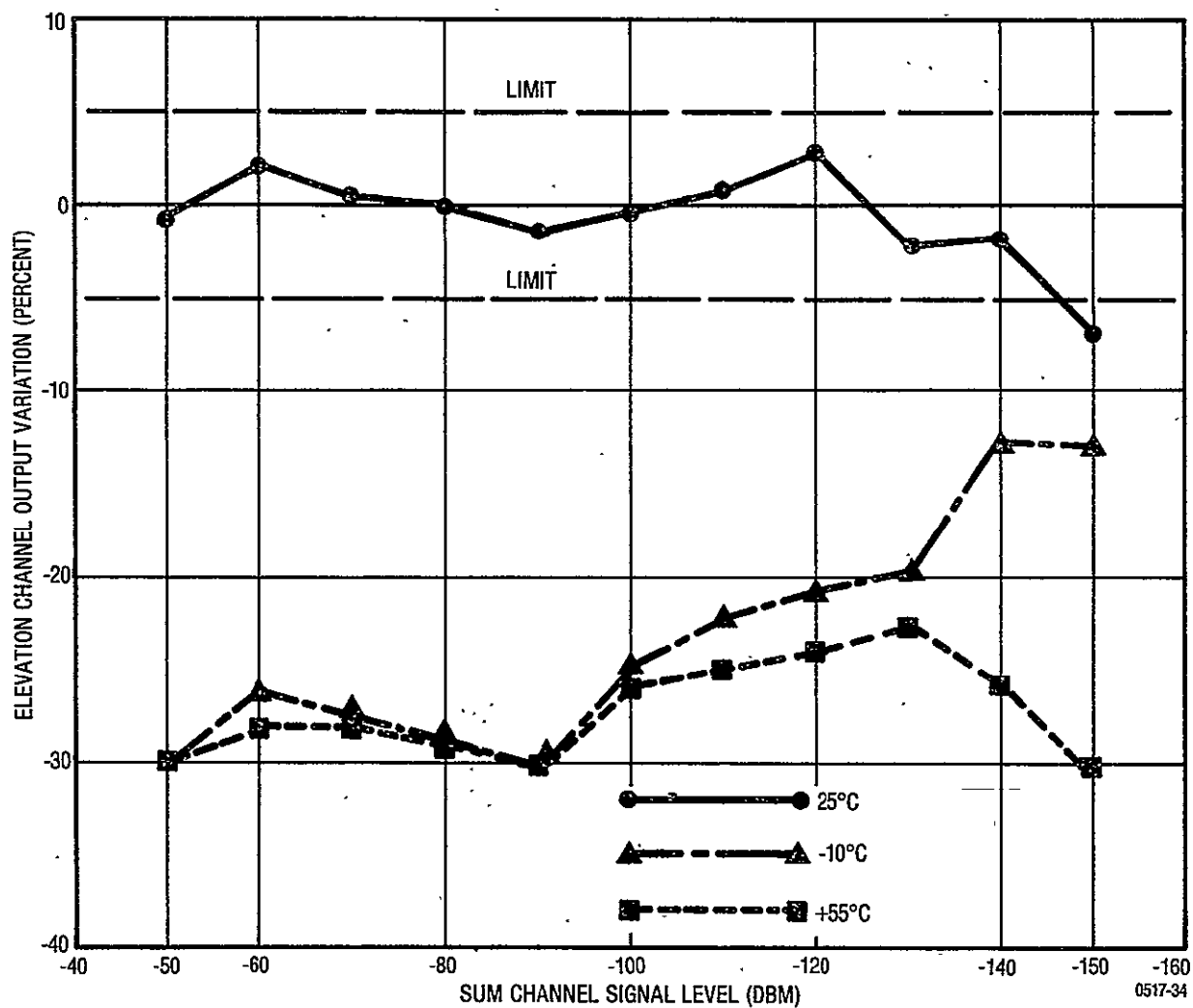


Figure 7-23. Quadraphase Elevation Channel Amplitude Response vs Signal Level and Temperature

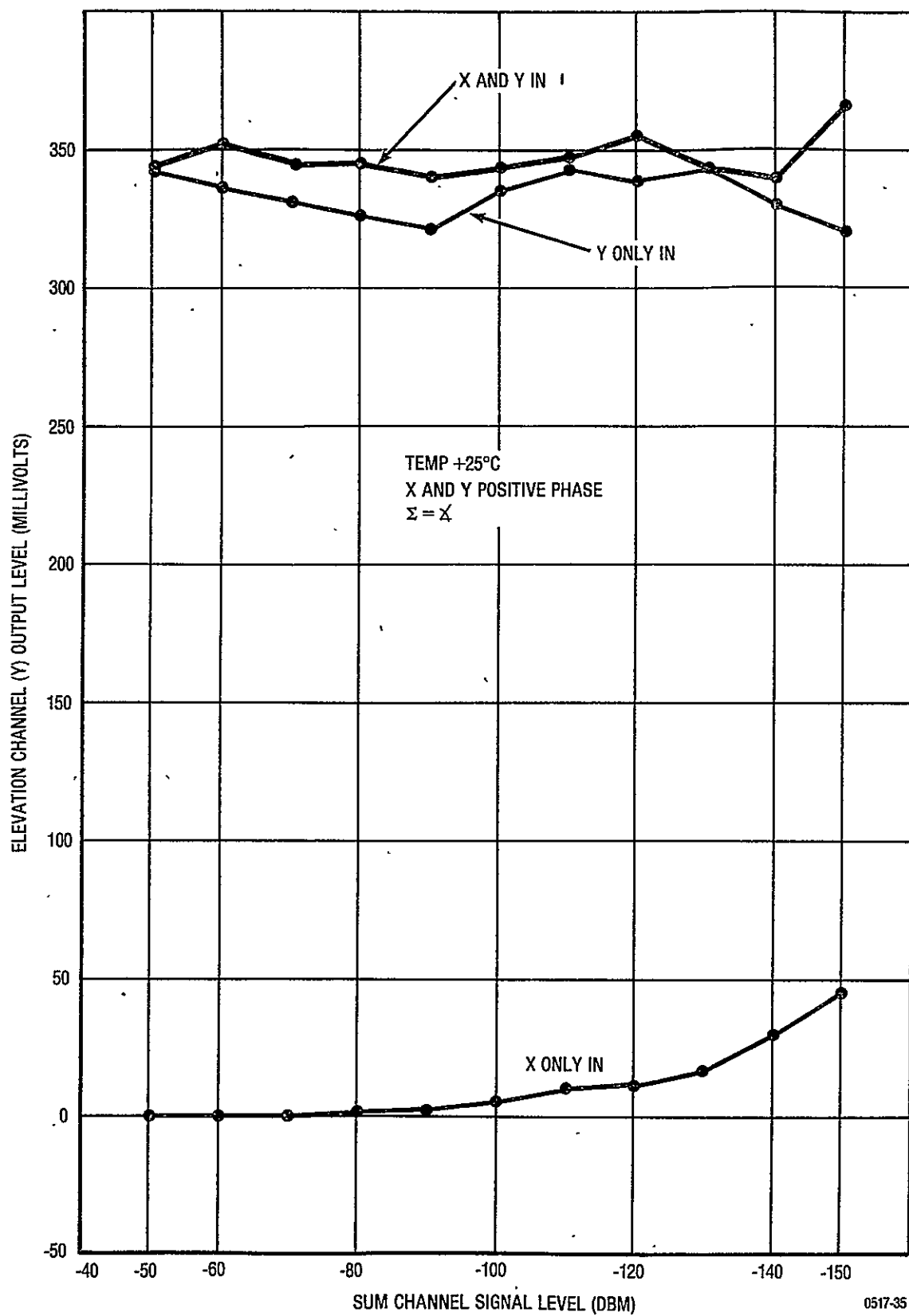
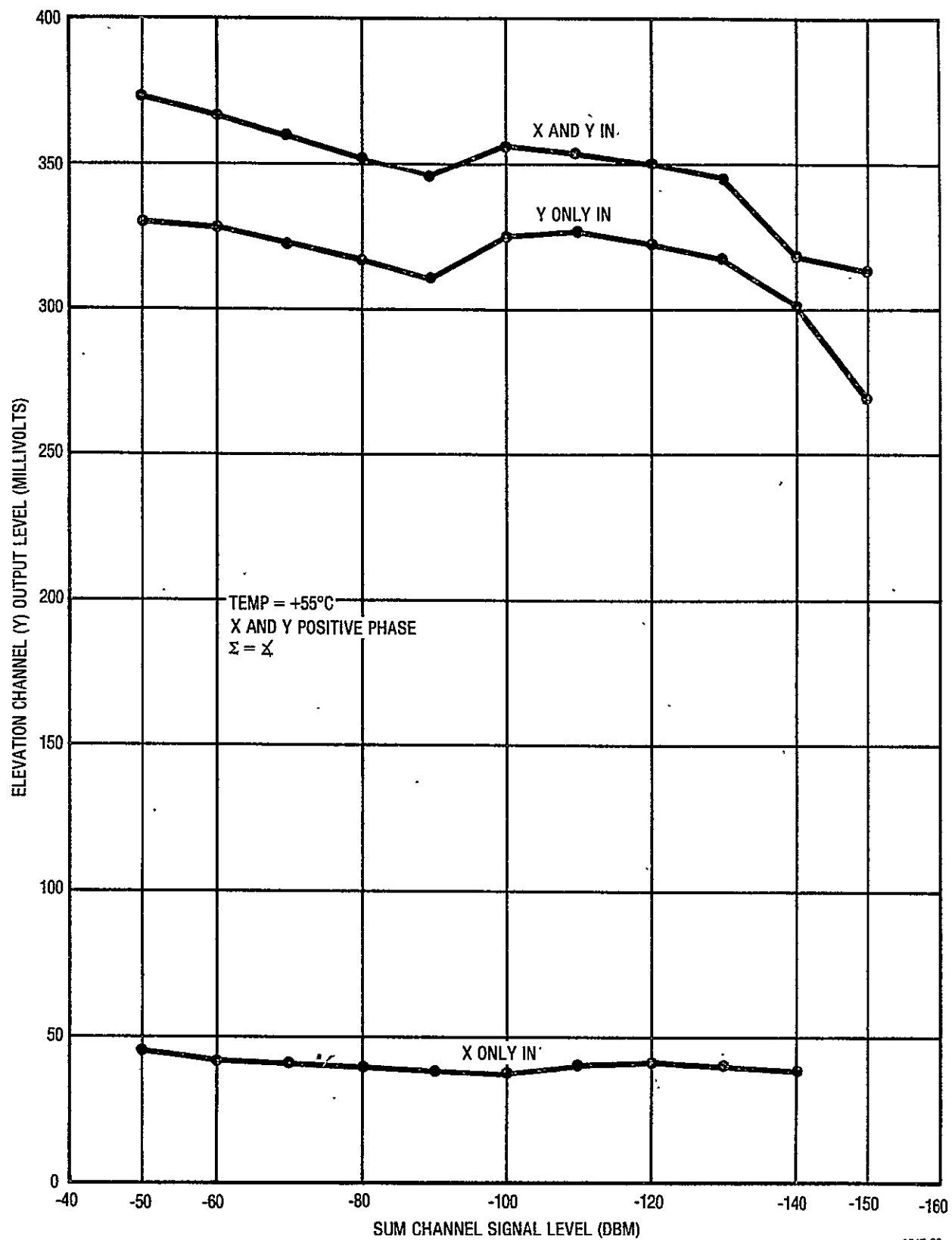


Figure 7-24. Quadrphase Elevation Channel Output Level vs Various Inputs



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Figure 7-25. Quadrphase System Elevation Channel Output Level vs Various Inputs

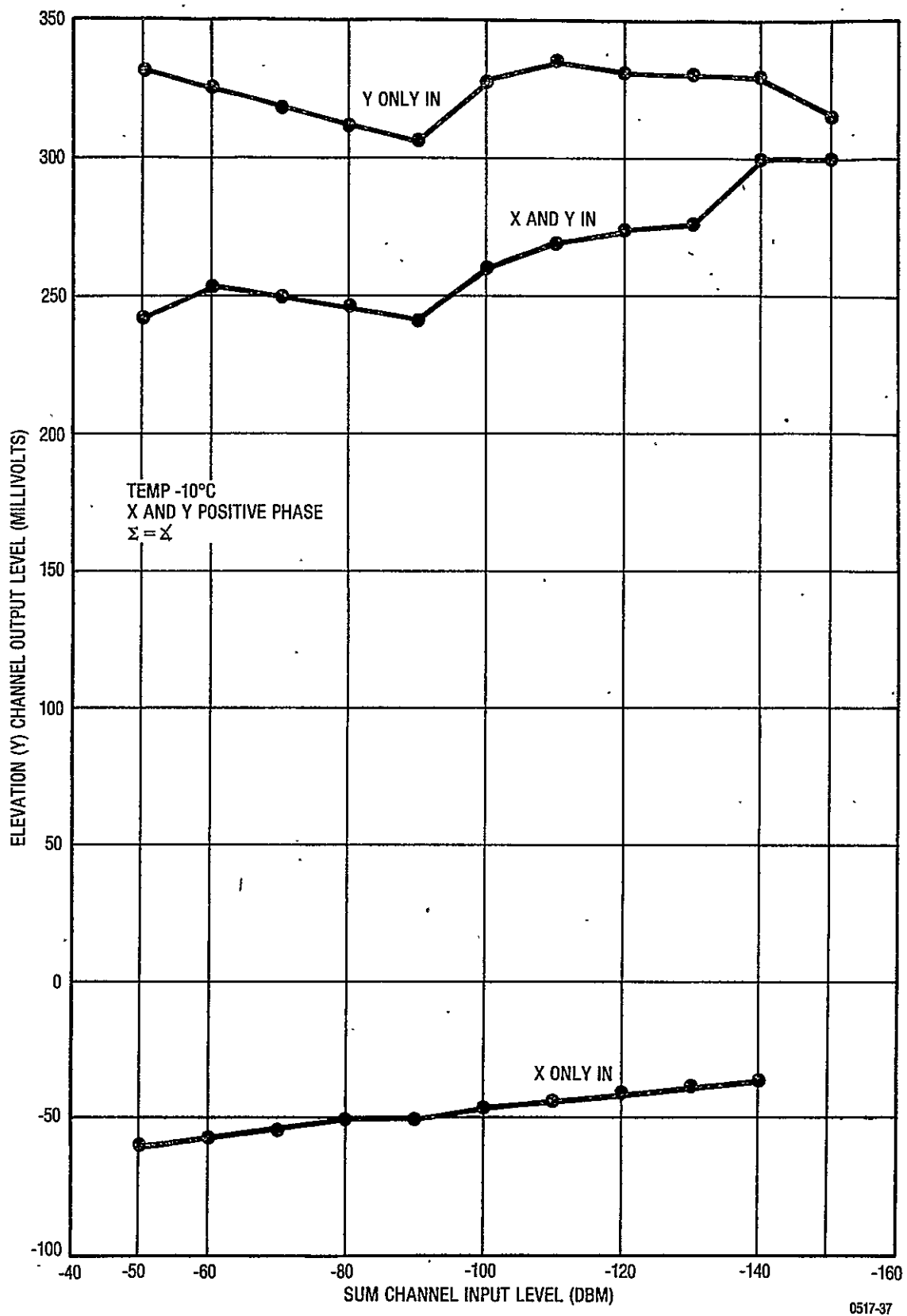


Figure 7-26. Quadraphase Elevation Channel Output Level vs Various Inputs

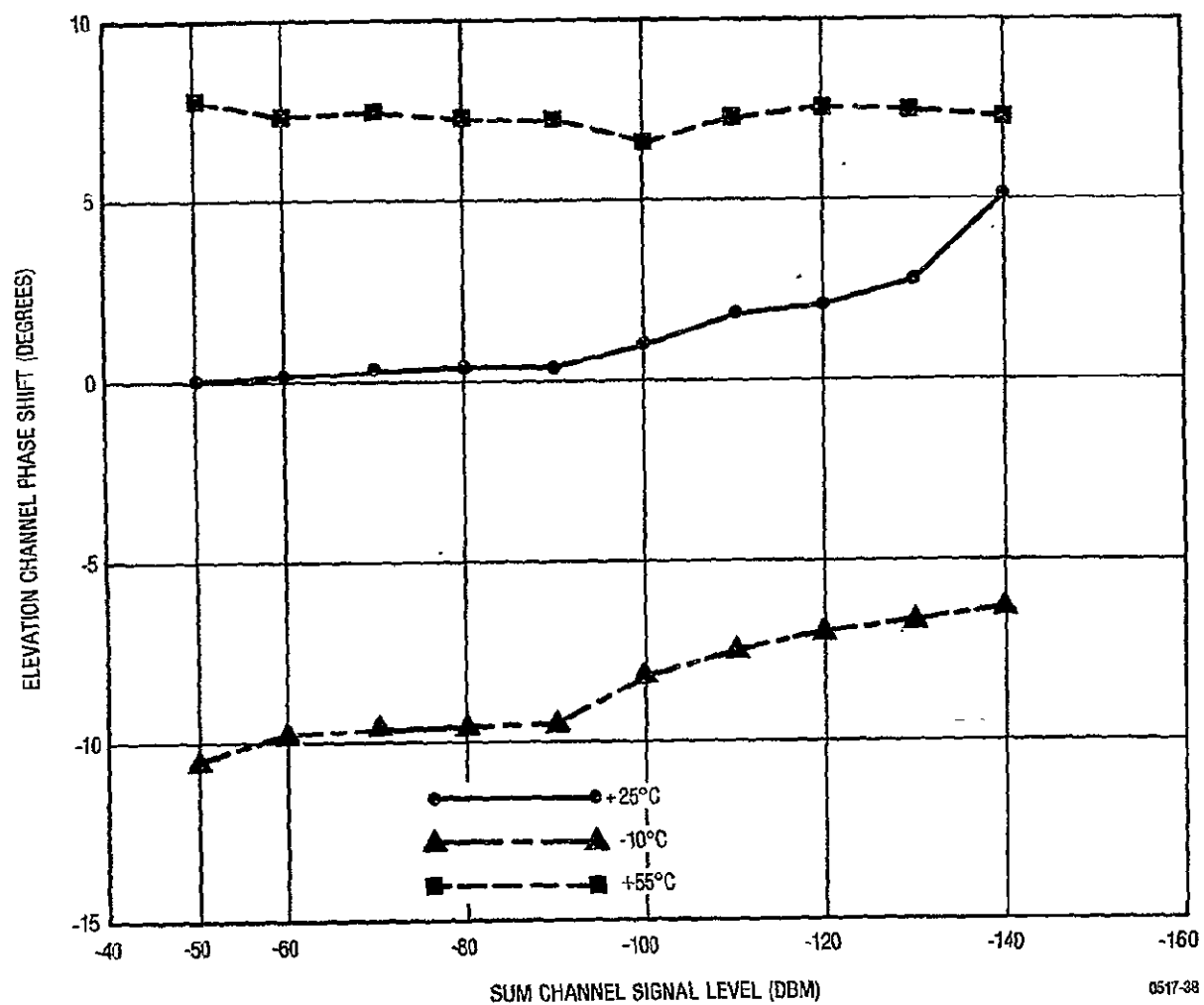


Figure 7-27. Quadraphase Elevation Channel Phase Shift vs Input Level and Temperature

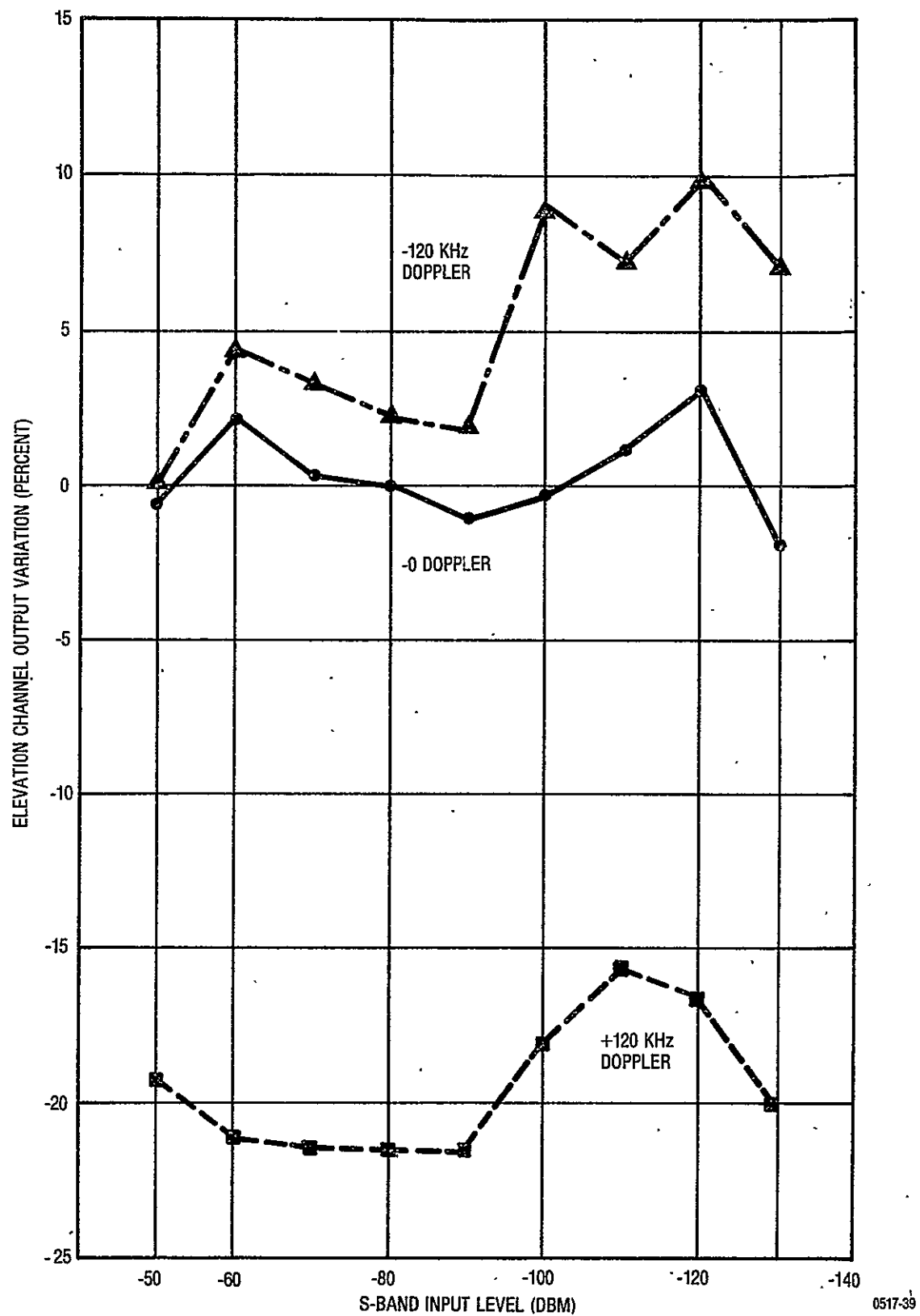


Figure 7-28. Quadrphase Elevation Channel Output Variation with Doppler and Temperature +25°C ($\Sigma = \Delta$)

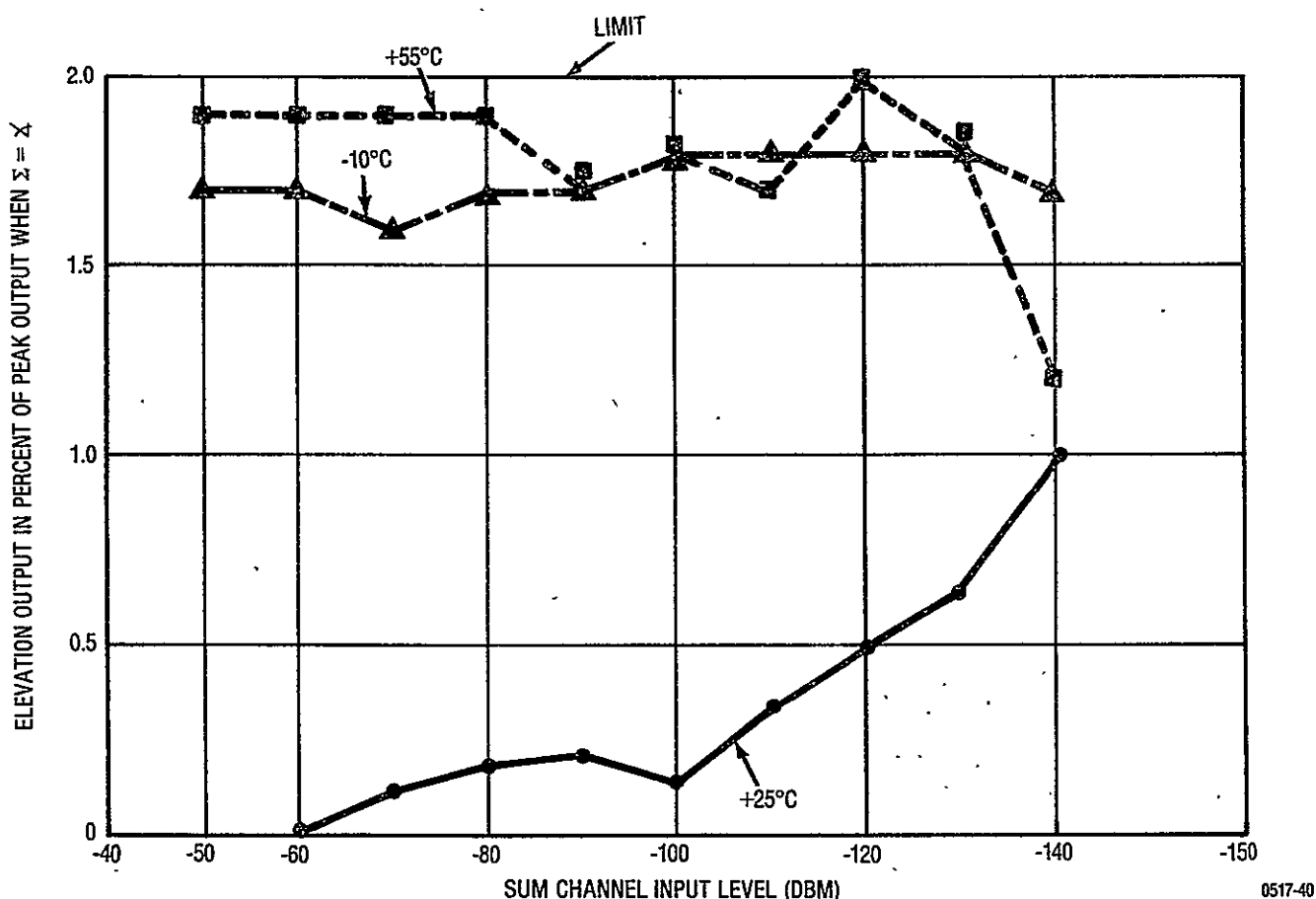
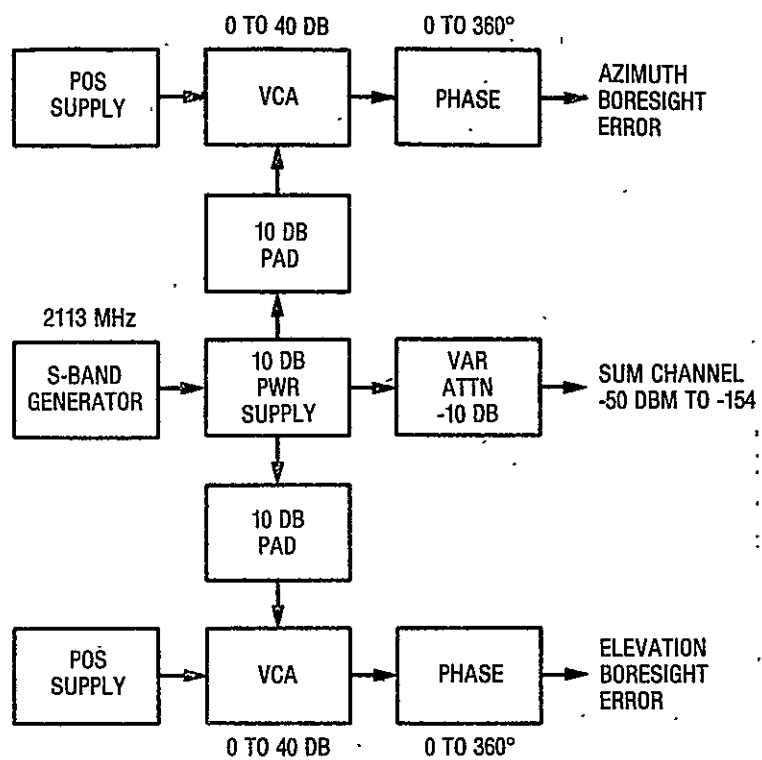


Figure 7-29. Quadraphase Elevation Channel Balance vs Signal Level and Temperature

7.3 Test Set

Figure 7-30 shows the S-Band signal generation scheme for providing sum and error signals to the monopulse receiver.

The voltage controlled attenuators (VCA's) in both the azimuth and elevation error channels were designed and fabricated by Motorola. All other pieces of RF hardware, (phase shifters, variable attenuator, power splitters, and 10-dB pad attenuators), are standard Hewlett-Packard RF test equipment. Each of the VCA's is controlled by a precision positive power supply for an attenuation ratio of 0 to 40 dB of the error channel S-Band signals. All S-Band signals are coherent with a single S-Band generator. Calibration of the error channel signal levels was accomplished by connecting each error channel output to the transponder (sum channel), and calibrating the respective VCA in volts per dB from the receiver's AGC calibration. The variable phase shifters were then calibrated by maximizing the respective monopulse receiver boresight error output for each attenuation level of the VCA's.



VCA ATTN	NULL DEPTH Σ/Δ RATIO	SYSTEM OUTPUT
0 DB	0	316 MV RMS
-10	10	100 MV RMS
-20	20	31.6 MV RMS
-30	30	10 MV RMS
-40	40	3.16 MV RMS

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Figure 7-30. S-Band Test Set

SECTION 8

8. CONCLUSIONS

Evaluation of the test results of the Time Division Multiplexing and Quadrature Systems indicates that the measured performance followed closely to that predicted by the initial analysis.

The major significant conclusion is that the TDM System, when operating with a NASA Standard Transponder, can meet the overall requirements specified in Section 3. However, this does require careful alignment and temperature compensation.

The Quadrature System displays relatively poor angle channel crosstalk performance which is due to the inherent differential phase tracking of the sum angle channels.

The TDM System displayed crosstalk capability of greater than -50 dB while the Quadrature System was capable of only -15 to -20 dB over the temperature range.

In addition, chopper stabilization of the detector/amplifier circuits was required to meet the dc balance requirements. This added a penalty to the TDM System since two separate detectors were required rather than just one. For the Quadrature System of course, two detectors are required. The TDM System also requires more power than the Quadrature System due to the requirement for an S-Band switch.

The transponder performance appeared not to be degraded with the addition of the angle channel. No measurable difference was detected in the ranging or command function signal-to-noise ratio's or degradation of system threshold performance. The best lock frequency did show a distinct difference with the angle channel energized, a factor which was not investigated further at this time. Fabrication of the angle circuitry into beam-lead modules presents no major problems since most of the discrete parts are generically the same, with the exception of the CMOS digital circuitry. However, it is felt that this is not a problem with the use of flatpacks mounted in a Standard Transponder module.

SECTION 9

9. RECOMMENDATIONS

The Microminiature Receiver which was used as the monopulse sum (reference) channel displayed several problems which should be investigated and corrected before further tests are conducted. First, it is highly probable that a $13F_1$ (248 MHz) coherent spur is present in the first LO chain at an approximate -175 to -180 dBm level. This does not affect the sum channel to an appreciable amount but does degrade angle channel performance at input levels below -140 dBm. Secondly, the sum channel displayed tracking problems with Doppler offsets which affected angle channel measurements.

An S-Band isolator was added in the sum channel first LO path to provide an additional 30 dB of isolation of the sum channel signal leakage to the angle. Selection of this isolator should be made on the basis of the requirement of at least 80 dB of total to angle first LO isolation.

Another area of further investigation is the tradeoff of guard bands requirements and S-Band switch switching speeds. The use of guard bands adds additional complexity to the timing logic; however it does relax the switch rise and fall time requirements proportionally, which should allow a major savings in power requirements.

A digital timing scheme is recommended to synthesize the control signals and guard times using a digital synthesis technique.

This report was not required to address the electrical and mechanical impact on the NASA Standard Transponder, but to show that such an antenna pointing system is demonstrable with a typical Standard Transponder with little or no impact on the transponder performance. However, it is felt that further study is required to fully evaluate the electrical and mechanical impacts on extracting the reference signals which are supplied to the error channel.